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Audio Out

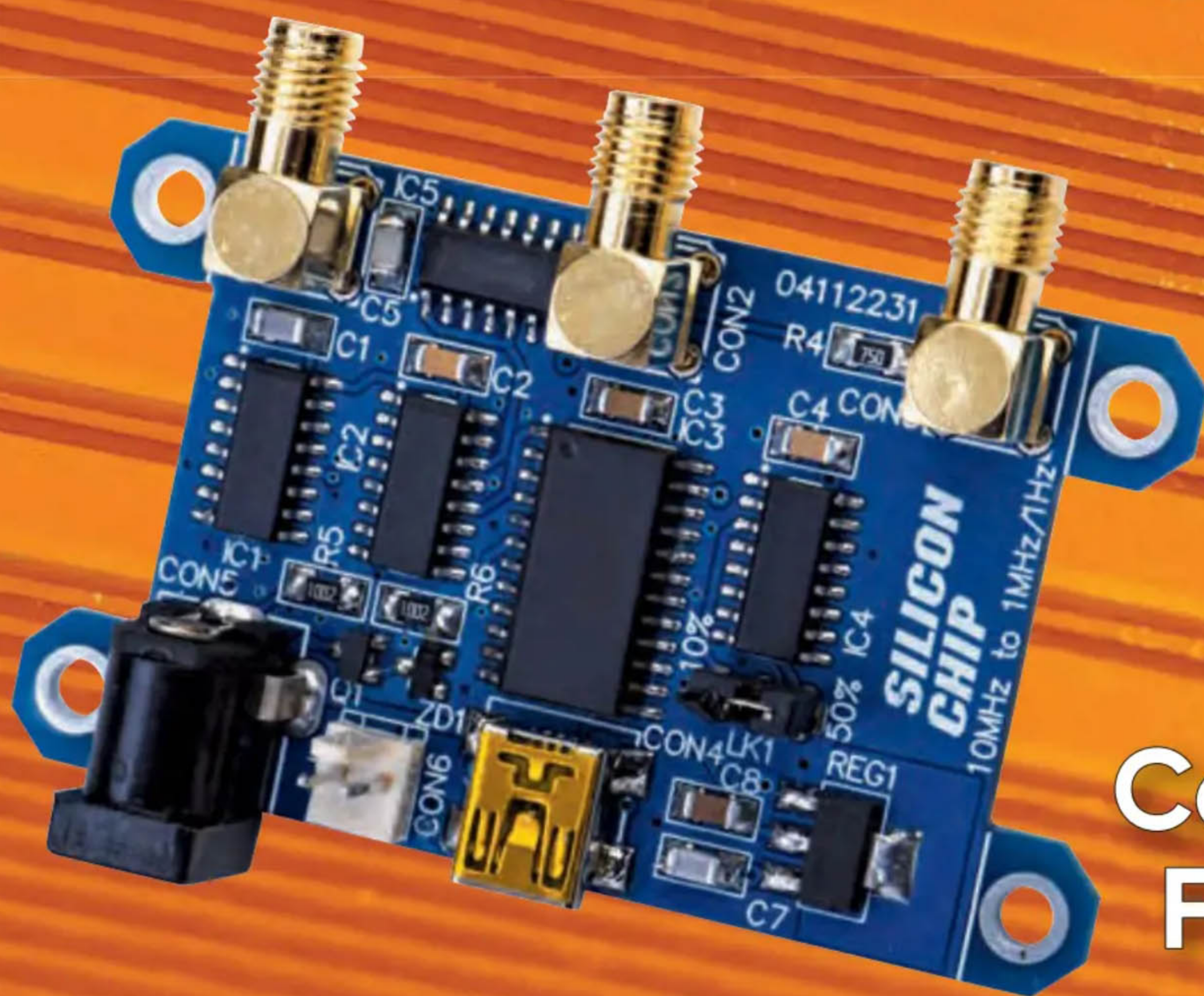
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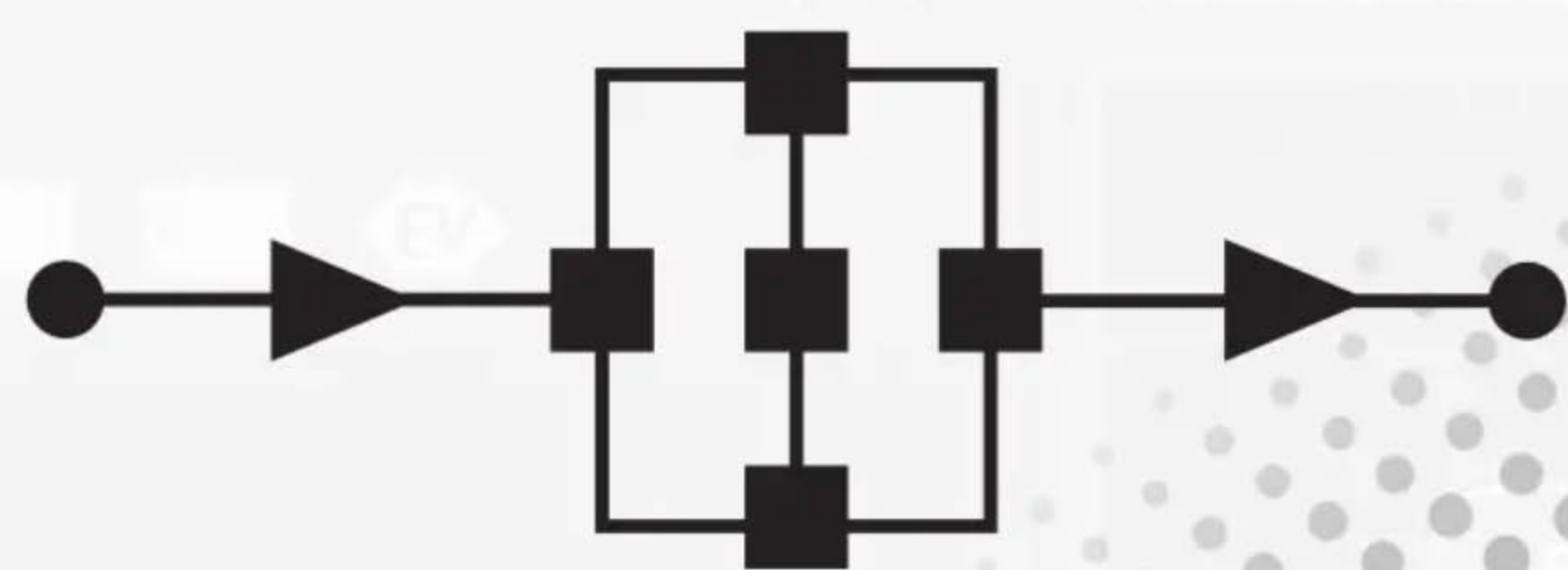
Techno Talk – AI-enabled holograms

Cool Beans – building a 14 × 10 tricolour LED array

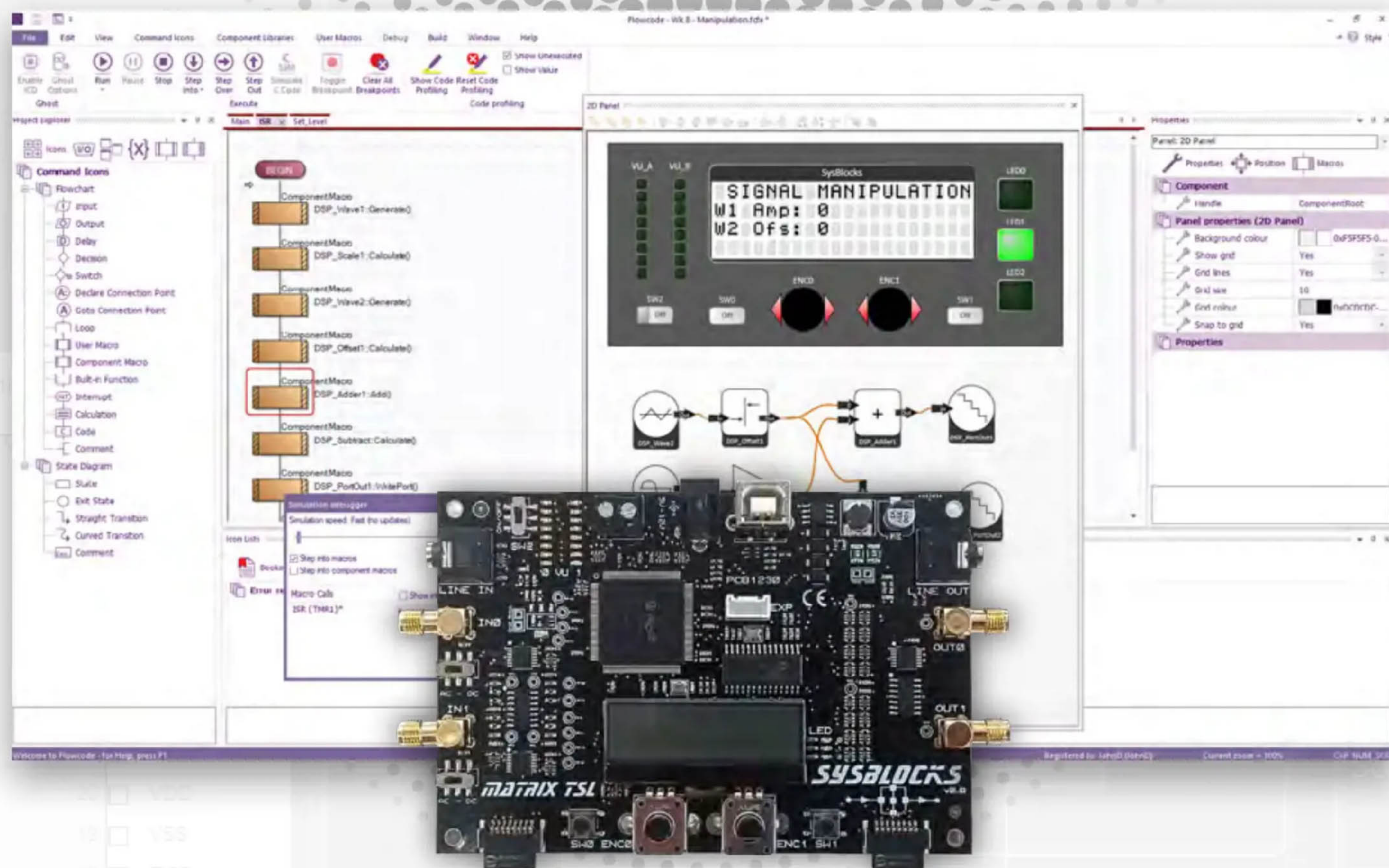
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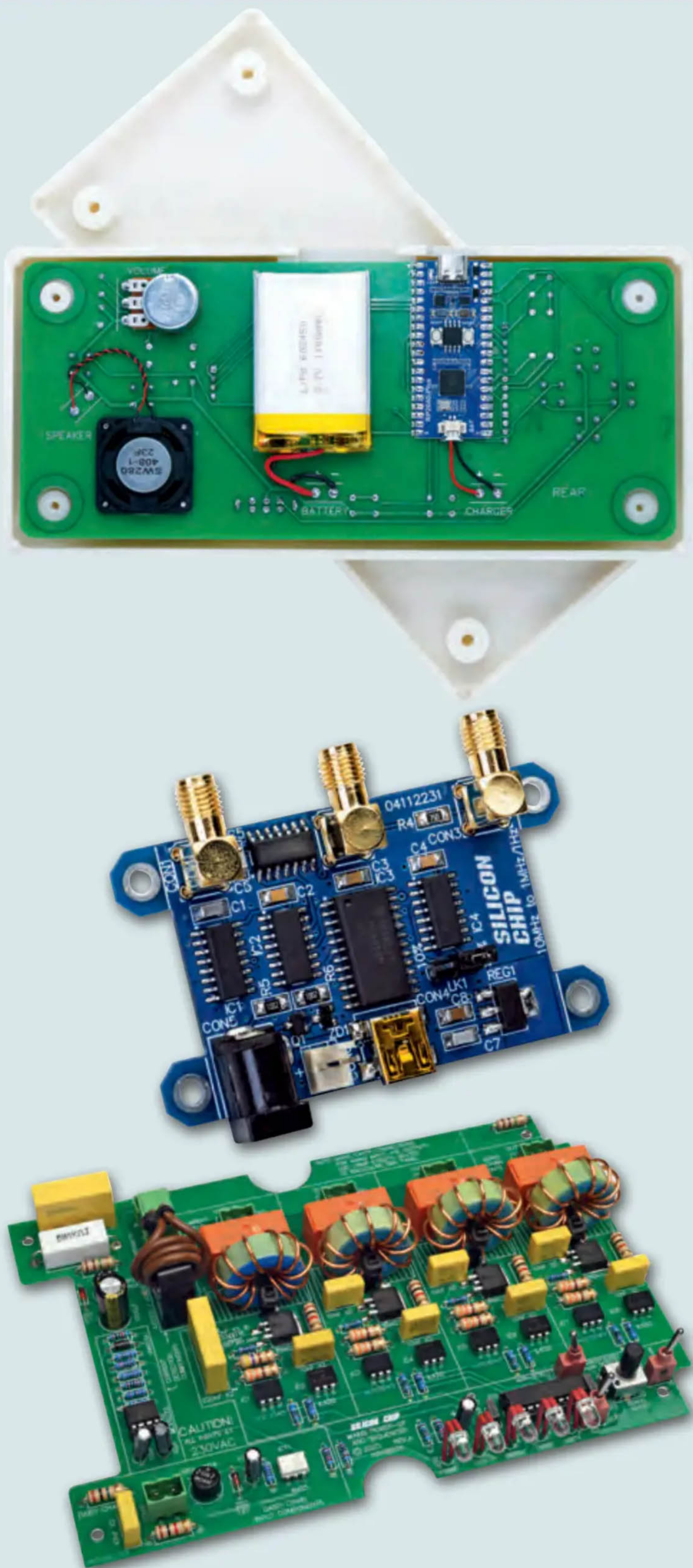


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Editorial offices

Electron Publishing (Aust.) Tel +61 2 9939 3295
Unit 1, 234 Harbord Road Email pe@pemag.au
Brookvale NSW 2100 Web www.electronpublishing.com
Australia

Address mail to:

Electron Publishing (Australia)
PO Box 194, Matraville NSW 2036
Australia

Advertising enquiries

+61 2 9939 3295 pe@pemag.au

Editor	Nicholas Vinen	
Publisher	Nicholas Vinen	
Digital subscriptions	Stewart Kearn	Tel 01202 880299
Online Editor	Alan Winstanley	
Web Systems	Kris Thain	
Production	Bao Smith	
Technical staff	Tim Blythman, John Clarke	

Print subscriptions

Practical Electronics Subscriptions
PO Box 6337
Bournemouth BH1 9EH Tel 01202 087631
United Kingdom Email pesubs@selectps.com

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Projects and circuits

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Some projects and circuits published in *Practical Electronics* employ voltages that can be lethal. Do not build, test, modify or fix any mains-powered equipment unless you fully understand the safety aspects involved and you use an RCD (GFCI) adaptor.

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Microsoft's constantly moving target

Because I like to feel that I own my computer, and I want to be free to do whatever I want with it, I run Linux at home. I've been using it for probably 20 years now and it certainly has come a long way. When you know how to use it, it is extremely capable.

Unfortunately, I can't really do that for my work computer. While there are other packages, most large publishers use Adobe products like InDesign, Illustrator, Photoshop etc. I also need to use Altium Designer for PCB design work and CorelDraw too. None of these programs are available for Linux, nor do they run well in emulation.

So basically my choices are macOS or Windows. Altium Designer only runs natively on Windows and CorelDraw runs very slowly on macOS, so Windows it is. I don't have a major problem with Windows 10; I feel that it is inferior to Linux in many ways (mainly in terms of performance and configurability) but I am used to it.

Of course, now that I'm used to it, Microsoft is threatening to end support for Windows 10. They're certainly trying to get people to move to Windows 11, but I've tried it, and it's miserable – definitely an overall worse experience than Windows 10.

The *Fox Report* in this issue goes over some of the trials and tribulations involved in 'upgrading'.

I've been forced to migrate several times thanks to Microsoft, from Windows 2000 to 7 to 10. Each time I felt like it was basically pointless and I got nothing out of it, other than not being 'left behind'. So I'm not keen to do that again.

Happily, I discovered the website <https://0patch.com>

This is a Slovenian company that provides security patches for Microsoft Windows. Unlike Microsoft's patches, apparently you don't have to reboot to install them. They say they will continue to support Windows 10 for at least five more years (I suspect it will be longer than that).

The service is free for personal, non-profit educational and testing use. Otherwise, it costs €24.95 (about £21) per computer per year. I don't know about you but I would gladly pay that to avoid dealing with any more of Microsoft's shenanigans and forced reboots for a few years.

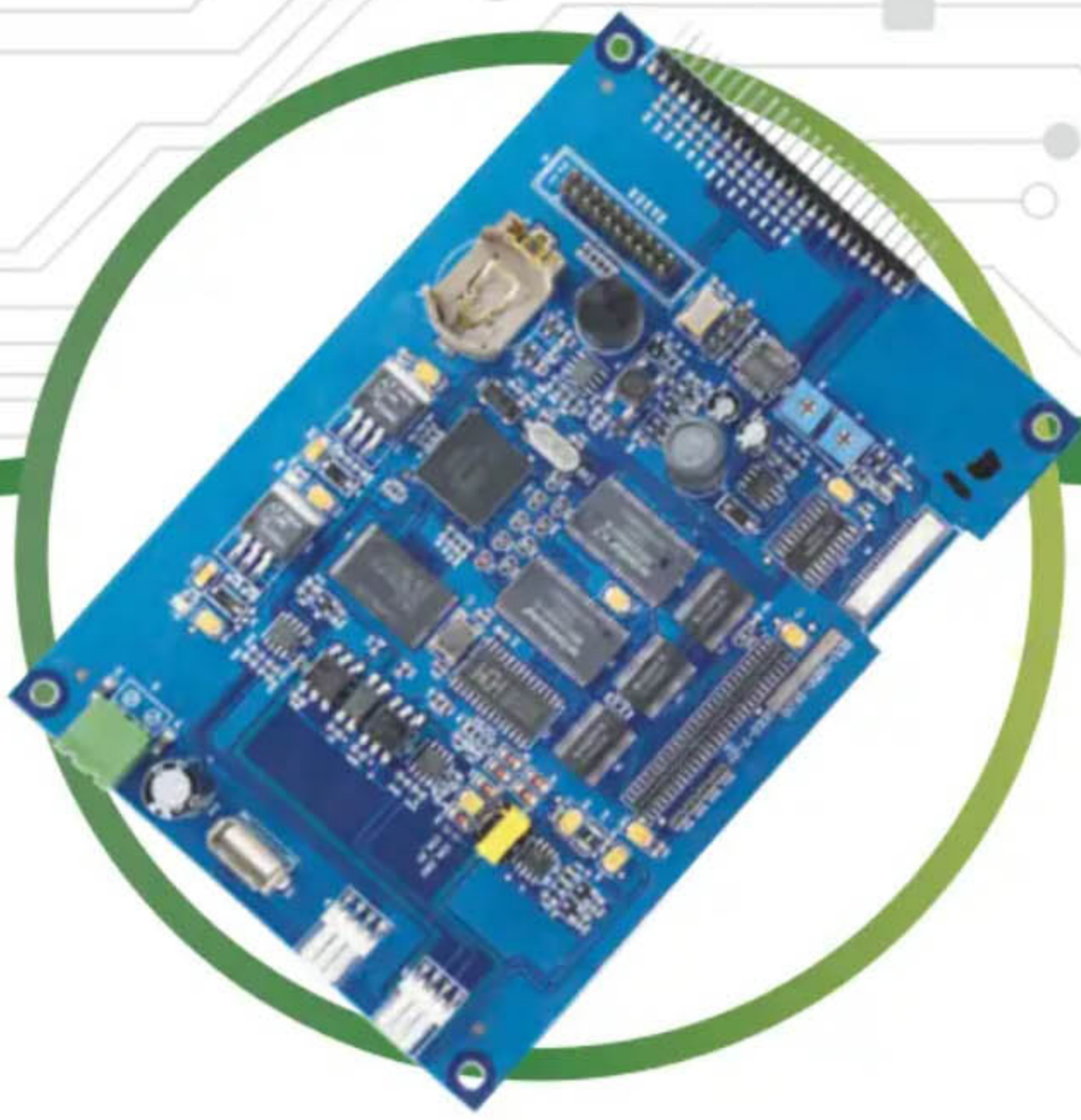
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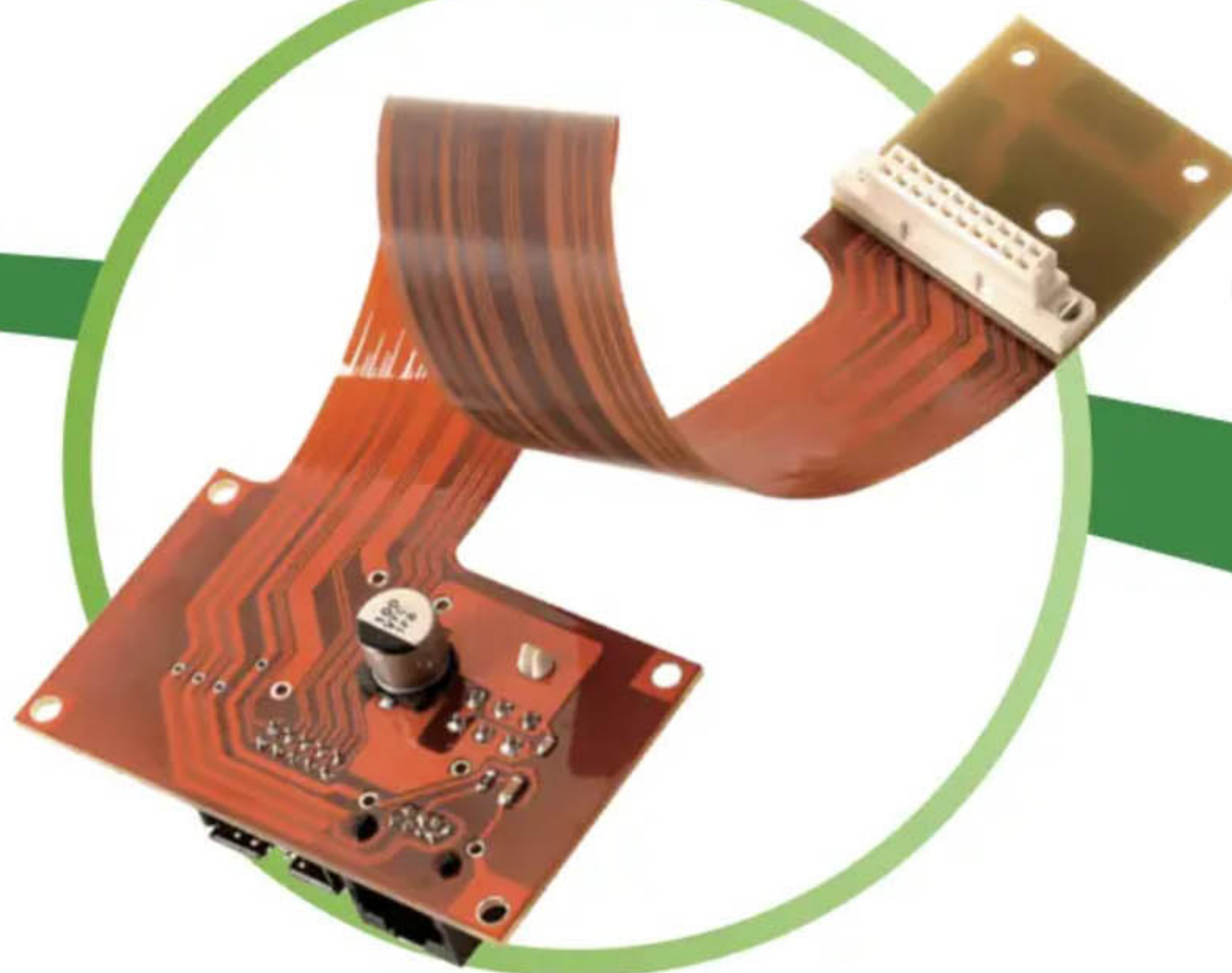
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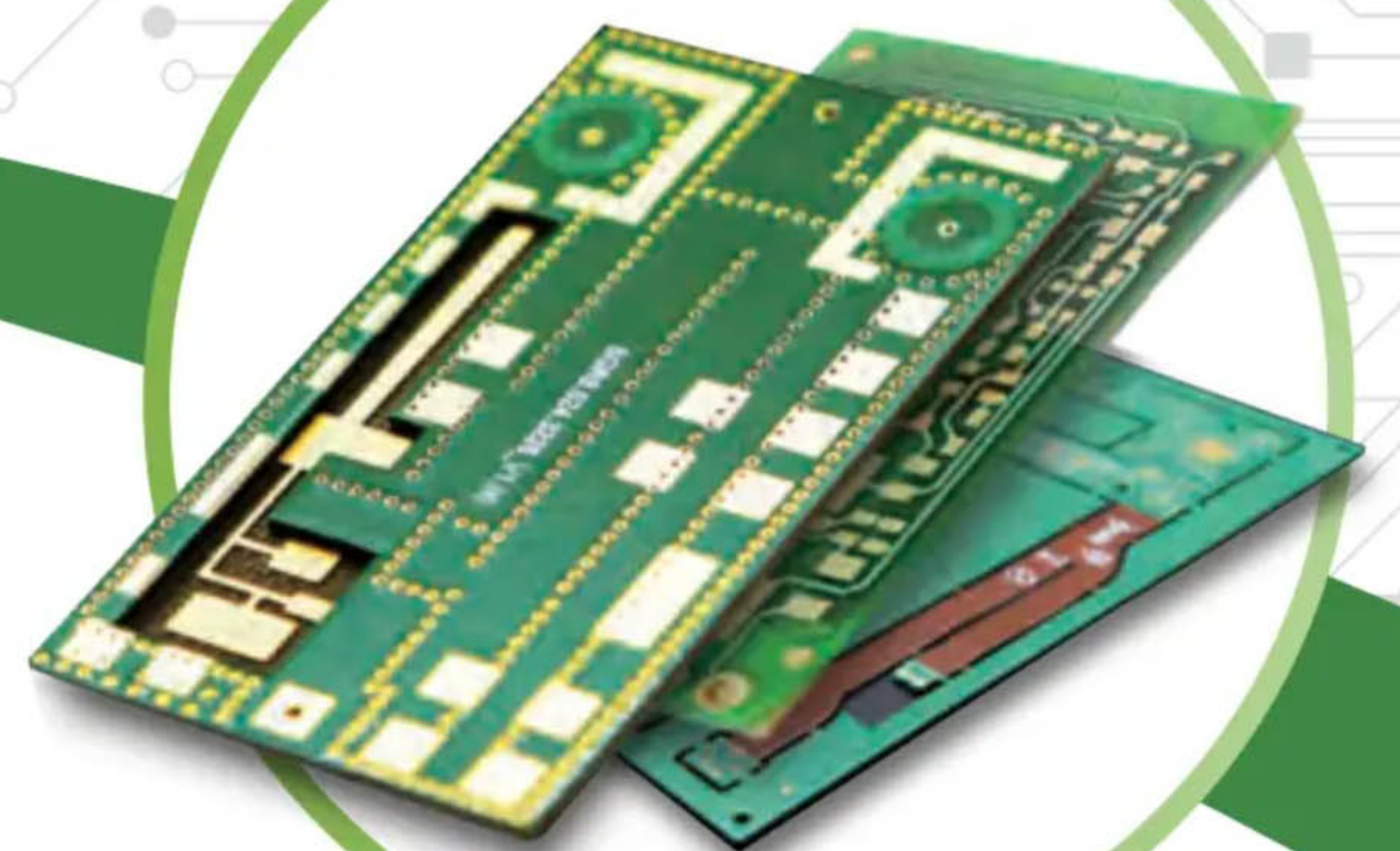
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Net Work

Alan Winstanley

This month's Net Work brings a round-up of the latest space program developments, starting with a look at some home-grown projects that are slowly taking shape in Britain and around the world.

As Europe seeks to build up its own independent capacity for launching satellites into low-Earth orbit (LEO), Scotland is in pole position to host some of Britain's own space port ambitions.

Back in the May 2022 issue, I reported that the UK Government was investing in SaxaVord, a new spaceport to be built in Unst, the most northernmost location in the Shetland Islands off the Scottish coast. Launches as early as 2024 were envisaged but they didn't materialise.

The spaceport is open for business as the UK's first vertical orbital site licensed to offer launch facilities to its clients. The site could handle up to 60% of all of Europe's LEO satellite launches, they claim, but the rockets themselves are still undergoing development. Successful engine tests have already been held, and the first launch may take place as soon as this year, it is hoped.

I also reported in 2022 how a second launch site was taking shape, in the form of Sutherland Spaceport, which aimed to become "Scotland's Sustainable Spaceport" in the Scottish Highlands.

Apart from anything, this project – the first carbon-neutral spaceport in the world – would bring some very welcome, high-tech jobs for young people in an area that is otherwise steeped in the traditions of crofting (a unique historical system of land tenure and small-holding farming).

Crofters are very passionate about managing the land handed down to them, but opportunities for employment are very limited, so hopes were high that a space hub could attract new sustainable jobs. Readers can gauge the crofting lifestyle for themselves at <https://www.melness.scot/news>

Sadly, despite its green credentials, the site's potential environmental impact has drawn much ire and opposition in some quarters. So much so that in December, Orbex, one of the spaceport's main backers, switched horses mid-race to focus on launching from SaxaVord instead.

They may return to Sutherland in the future, they say, but the decision enables Orbex to focus instead on its new, medium-sized launch vehicle called Proxima. The company plans to enhance its manufacturing facilities

in Forres to support the production of both its smaller vehicle, known as Prime, and the new Proxima launcher.

Skyrora (<https://skyrora.com>) is a rival currently designing its own satellite space launch vehicles, including complex 3D-printed assemblies, ready to meet the needs of its LEO satellite clients. They forecast flights commencing at the end of 2026.

England has had its fair share of excitement in the emerging space program, with Spaceport Cornwall having been the venue of the ill-fated Virgin Orbit mission to fly the Cosmic Girl 747 in January 2023, lifting the Launcher One satellite carrier into low orbit. The plane still flies today as the Spirit of Mojave.

While budgets are just a fraction of, say, NASA's or SpaceX's, there is no denying the drive and eagerness to build an independent space launching system in the UK. During the next year, we can finally expect to see the first LEO satellites being launched from British soil.

Jostling for space

Vying for business is Sweden's Esrange Spaceport, which is home



SaxaVord Spaceport is developing a ground station network on the island of Unst, Shetland, the UK's most northerly location.

to a thriving and long-established program of scientific and research initiatives. Esrange is the European Union's first site capable of offering orbital launches. You can learn more at <https://sscspace.com/esrange/>

Europe's main spaceport is situated in French Guiana on South America's northern coast. Its location near to the equator means rockets gain extra performance, thanks to the impetus of a 'slingshot effect' due to the earth's higher rotational speed, as the ESA puts it.

The ESA is focusing on two space vehicles that will doubtless become familiar news items in the years ahead: the smaller Vega-C has returned to action after being grounded for two years, and it successfully launched an environmental monitoring satellite last December.

The ESA's much delayed Ariane 6 heavy lifter is scheduled to launch in the first quarter of 2025. Two versions of the space vehicle are offered: the twin-booster Ariane 62 or four-booster Ariane 64, the latter capable of lifting up to 21 tonnes into LEO.

The vehicles are manufactured in various plants across Europe, including facilities in France, Germany and Italy, before being shipped to French Guiana on board a new, purpose-built ocean-going vessel: the Canopée.

The vessel is equipped with four Oceanwings wind-assisted propulsion 'sails' that reduce the fuel consumption by 30% over a normal diesel-powered ship. They say they halve the costs of shipping Ariane 6 parts to the spaceport. Readers can expect to hear more of Ariane 6's missions in the next few years.

NASA's Artemis lunar program has been beset by delays and setbacks. Artemis is the mythological twin sister of Apollo and represents NASA's ambition to return humans to the moon and then enable the human exploration of Mars.

An unmanned mission in 2022 eventually saw an Orion capsule fly past the moon and return successfully. The Artemis 2 mission will carry astronauts the furthest humans have ever travelled into deep space, but it has been delayed to at least 2026.

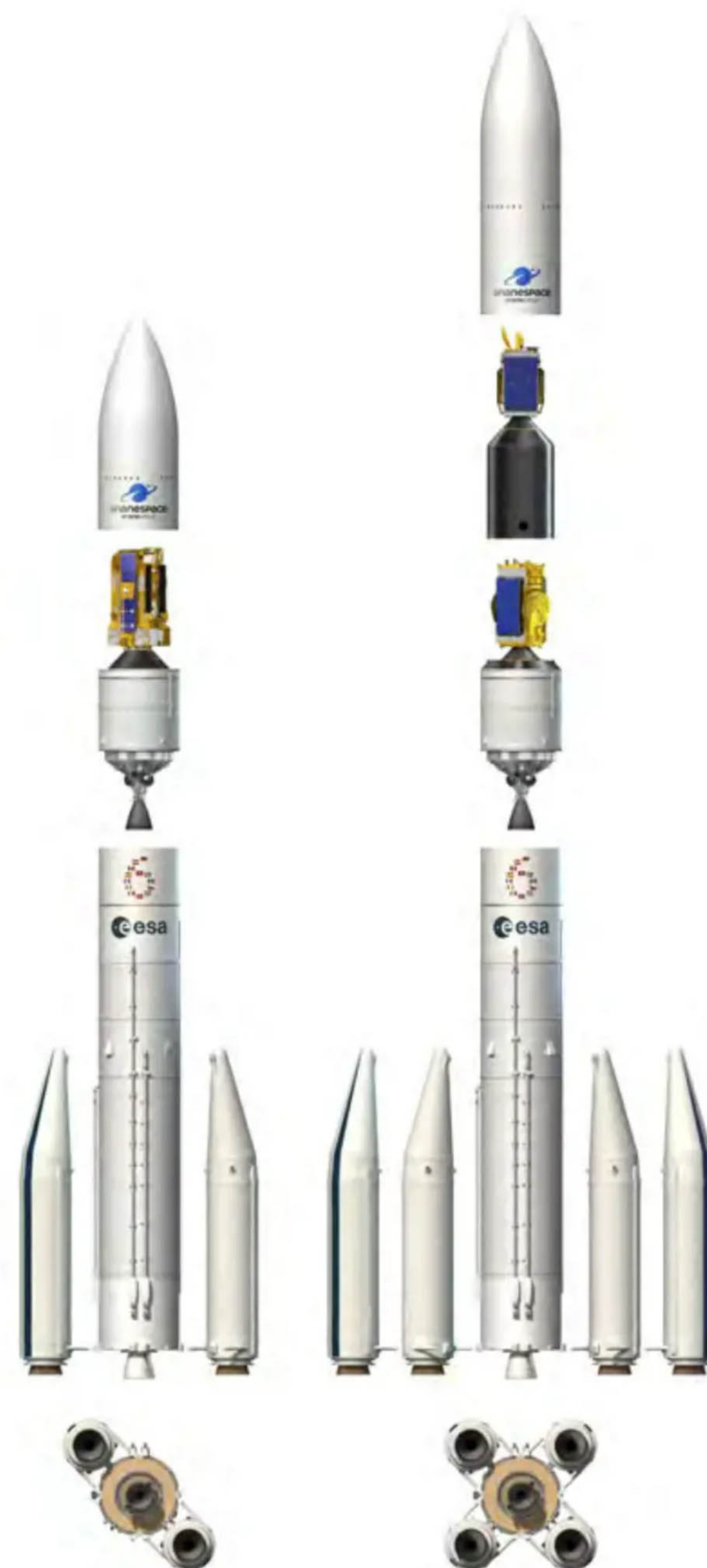
Sojourn on Starliner

Boeing's Starliner is designed to transport crew to and from the International Space Station. After several false starts, in June 2024, a crewed flight docked with the ISS on supposedly an eight-day mission. Question marks over the capsule's thrusters meant NASA delayed the astronauts' return to Earth.

Initially, the delay was for a few days, then a few weeks. Eight months later, the two astronauts remain stranded on board the ISS. The Starliner capsule returned to Earth unladen, and the two astronauts are finally expected to come home on a SpaceX Dragon capsule during the early part of this year.

Back in 2021, NASA's Mars Ingenuity helicopter made history after being the first vehicle to make a powered flight on another world. The proof-of-concept machine was expected to manage a handful of flights in the Martian atmosphere, but it proved so successful that the plucky Ingenuity flew more than 70 missions in total.

In April 2024, it was finally grounded due to rotor damage, but the electronics



The ESA Ariane 6 heavy lifter is set to launch from French Guiana early this year. It has two booster configurations depending on payload requirements. Source: ESA ArianeGroup.



The ESA Canopée hybrid transporter vessel carries Ariane 6 from mainland Europe to French Guiana. Special 'sails' reduce its fuel consumption. Source: ESA ArianeGroup, Tom van Oossanen.

will continue to capture data on board in the hope that it may be recovered by a future mission.

One NASA project that has needed a major rethink is the Mars Sample Return mission. As described in earlier *Net Work* columns, NASA's Perseverance rover had gathered an array of small sample tubes of Martian soil, each carefully logged and deposited onto the surface ready to be collected some time in the future.

Spiralling costs have put paid to an over-elaborate, multi-national mission that would see the sample tubes fetched into Martian orbit for eventual return to Earth. Lower-cost options are now being sought. China is also thought to be investing in its own sample return mission and may beat NASA to it, albeit in a much more haphazard 'smash and grab' sort of way.

Solar sailor

NASA's Parker Solar Probe, launched in August 2018, was the first spacecraft to pass through the Sun's corona. In December, it set another record for passing the closest to the Sun's surface at 430,000 miles per hour (692,000km/h), faster than any man-made object has ever travelled. It came within 3.8 million miles (6.1 million kilometres) of the Sun's surface.

Remarkably, all instruments and telemetry are working normally after the perilous flight.

SpaceX continues to expand its Starlink satellite network in LEO with a further 21 satellites launched in early January. The total number of Starlink satellites in orbit is fast approaching 7000, and astronomers and scientists are increasingly concerned that radio noise is interfering with their highly sensitive instruments.

With a can-do spirit, Elon Musk's SpaceX is also pressing on with developing Starship, the world's most powerful rocket that is ultimately destined for building a space camp on Mars. The re-usable rocket is powered by 33 Raptor engines and has previously taken off and landed again on the same launchpad.

In an astonishing achievement, the returning craft was captured and held aloft by 'chopsticks' on the launch tower. The video (<https://youtu.be/b28zbsnk-48>) must be seen to be believed. After five data-gathering test missions, you can expect the inaugural flight of Starship this year, followed by several other significant flights.

Blue Origin, the space technology firm owned by Amazon's Jeff Bezos, is also close to inaugurating its own heavy lift space vehicle, the New Glenn



A Starship super heavy booster with 33 Raptor engines is captured by 'chopsticks' after returning to Earth. Source: SpaceX (via YouTube).

rocket, named after John Glenn, the first American to orbit Earth.

Space tourism is also becoming a reality for the well-heeled, with passengers in the New Shepard capsule able to view the earth and experience weightlessness for a few moments. Amazon's Kuiper satellite network is also continuing to evolve, and will become a contender for offering space-based broadband.

As readers can see, the race to explore space is heating up. Reusable rockets can now be guided back to earth with pinpoint accuracy, even landing vertically on a carrier ship or the original launch tower. Several countries are developing remarkable technologies to explore the far side of the moon, the Sun and to bring samples back to earth.

Launching satellites into LEO no longer needs the resources of Cape Canaveral. Tiny cubesats, designed to carry experiments into space, measure just 10×10×10cm (about 4×4×4 inches) and fit in the palm of your hand. It's now routine to hitch a lift on a ride-sharing space mission where they can be released into a geostationary orbit.

At the other extreme, the largest rockets the world has ever seen are being built in preparation for the day when we may finally see humans landing on Mars.

eBay sellers beware

There was a time, say 20 years ago or more, when eBay was the go-to website to buy or bid for all manner



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of merchandise. Mostly they were private sales, but businesses were quick to jump on board and sometimes ‘game’ the system.

Online auctions could be a source of much excitement and in the era of slow dial-up internet or ADSL and no Wi-Fi, bidders were often outbid in the closing few seconds due to ‘sniping’ by other bidders.

Websites such as Auctionstealer and Bidnapper sprang up to enter last-second bids on your behalf. It was quite an engaging, if sometimes disappointing, process and for much of the time, eBay was fun to use.

With ever more choice from other sites chipping away at its core business, eBay has been re-inventing itself, and in January announced a new fee structure. Never before have buyers been forced to pay a fee on top (unlike, say, users of the Vinted website), but by the time you read this, eBay will have introduced its so-called “Buyer Protection” scheme.

The idea sounds grand, but, taking a leaf straight out of Vinted’s book, it simply increases the buyer’s costs by adding an enigmatic 4% “Buyer Protection” fee plus £0.75 when they purchase from private sellers.

Effectively, buyers are paying for their own insurance policy, but what do they get for their money? As a supposed added-value benefit, a 24x7 phone or chat service is introduced by eBay “to get the quick answers you need”.

Far more relevant to private sellers, though, is the fact that they will no longer be paid until the goods have been delivered successfully. This is supposed to “give buyers more confidence”, which they are supposed to be thankful for.

As any seller knows, eBay also tries to muscle in on the postage phase as well, by pushing its own shipping options for Royal Mail and courier deliveries. I doubt eBay does this out of charity, and I prefer to make my own arrangements anyway.

eBay’s actions are likely to have a chilling effect on small private sale items, which had only recently become free to list. Something that cost £5 will now cost £5.95 from February, as one disgruntled seller pointed out, making it much less viable to sell small items as costs are pushed onto buyers.

The onus is also now on the seller to provide tracking and deal with the consequences of deliveries going astray.

Should things go wrong, it’s worth remembering that buying through PayPal and Google Pay, instead of directly with your credit card, removes the Section 75 UK consumer protection law. Instead, you’re left to complain and deal with corporates like PayPal or eBay. In my view, these forms of payment are only suitable for low-cost items.

eBay’s woolly marketing gets worse: as the fee is already

included in the item price charged by private sellers, “there are no surprises at checkout”, they say, so “[buyers] only pay what you see”, which is jolly reassuring to know. Since there are no added fees for business sellers as “Buyer Protection” is already included, sales are likely to be skewed in favour of business sellers.

The exciting, auctioneering spirit of eBay is being all but snuffed out as it becomes a simple marketplace machine instead.

That’s all for this month’s feature. Next month, change is in the air, as the former Editor once put it. The April 2025 issue will sadly mark the last regular *Net Work* column after an uninterrupted run of nearly 29 years. So join me then for more news, and more *Net Work*!

The author can be reached at alan@epemag.net

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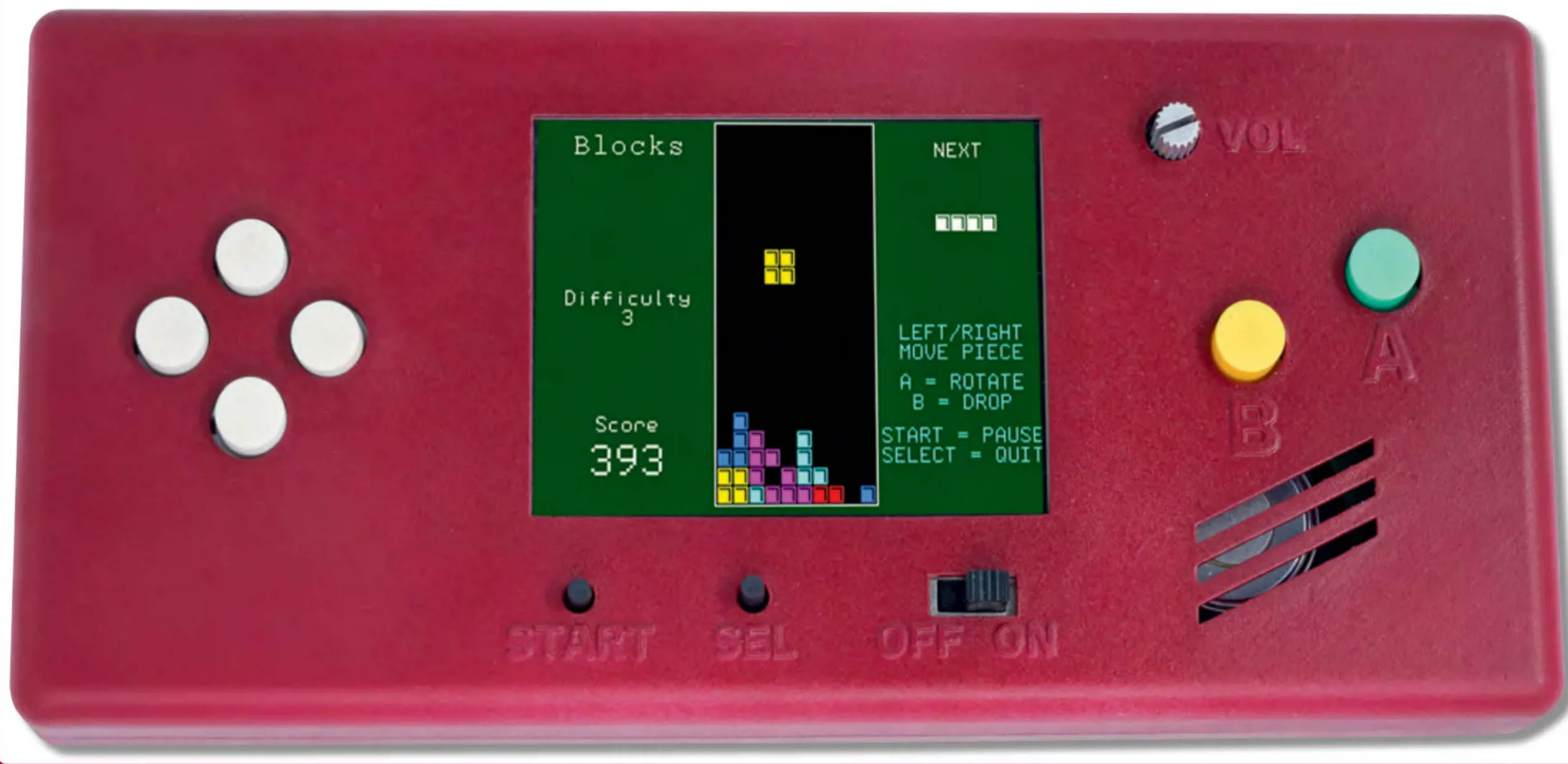
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Constructional Project

Project by Geoff Graham

THE PICO GAMER



The Pico Gamer is a PicoMite powered 'retro' game console packed with nine games including three inspired by Pac-Man, Space Invaders and Tetris. With its inbuilt rechargeable battery and colour 3.2-inch (81mm) diagonal LCD screen, it will keep you entertained for many hours.

The Pico Gamer was inspired by the Game Boy series from Nintendo, introduced in 1989. They were small handheld battery-powered devices, initially with tiny monochrome screens and an eight-bit CPU. Over time, more feature-rich versions were introduced, and the series became a massive success, with over 100 million sold across all variants.

The Pico Gamer is a marked upgrade on the original Game Boy, with a colour LCD screen, a dual-core 252MHz 32-bit processor, 2.5MB of internal game storage and a USB interface. It is easy to build, using just a handful of components, and fits nicely into a custom 3D-printed case.

We based our design on the layout of the Game Boy Advance, with the control buttons on either side of the screen in a horizontal layout. Such a design is a natural fit for a handheld game console and has since been adopted by many other consoles.

There are eight buttons on the console: four direction buttons (up, down, left & right) on the left, two control but-

tons (start and select) under the screen and two auxiliary buttons (A and B) on the right.

An important feature is the built-in rechargeable lithium-ion battery; the original Game Boy used four AA cells. The Pico Gamer's battery can last over eight hours, which is plenty for a long road trip, and can be recharged in under four hours using the built-in USB connector. It could even be recharged from a portable USB power pack in a pinch.

The 3.2-inch LCD screen has a 66×50mm active area containing 320×240 pixels. It can display over 65,000 colours, which most games use to good effect. The screen is also touch-sensitive, although currently, no games use that feature. New games can be programmed in BASIC, so perhaps one of our readers will come up with one that does!

The Pico Gamer has a mono audio amplifier and speaker for sound effects. Most games use them to create various beeps, squeaks and explosions. However, it is good enough to repro-

duce more complex sound effects and music, and some games do both. To see a gameplay video, visit:

<https://pemag.au/Videos/PicoGamer>

The processor

The Pico Gamer is powered by the Waveshare RP2040-Plus. This is a pin-for-pin compatible clone of the Raspberry Pi Pico, with a few important upgrades.

Firstly, it includes a lithium-ion battery charger, so that's one less feature that needs to be designed into the circuit. It also uses a high-speed flash memory chip, so the RP2040 processor can be reliably overclocked to 252MHz, which is required for the more processor-intensive games.

Finally, the RP2040-Plus has a 4MB flash memory chip rather than the 2MB of the standard Pico. This is important because we store the games in this flash memory, and the standard Pico has space for a limited number of games. However, the RP2040-Plus with 4MB can fit dozens of games, and you will be unlikely ever to fill that up.

The Pico Gamer

Features & Specifications

- » Dimensions: 198 × 90 × 22mm
- » Weight: 300g
- » Battery: internal 1100mAh LiPo battery
- » Runtime: approximately eight hours
- » External power/charging: 5V via USB at 260mA
- » Display: 66 × 50mm LCD, 320 × 240 pixels, 65,535 colours
- » Audio: 340mW from a 28mm diameter speaker
- » Internal storage: 14.5MB★ (sufficient for hundreds of games)
- » External storage: SD card up to 32GB, formatted as FAT16 or FAT32
- » Sound test and demonstration
- » Button test
- » File Browser
- » A selection of pre-installed games (see the panel opposite)

★ or 2.5MB if 4MB RP2040-Plus is used, sufficient for 30+ games

Using a standard Raspberry Pi Pico in the Pico Gamer would be possible. While it would work, you would have to keep it tethered via a USB cable for power. Also, you will need to install a custom version of the software because the software we supply is optimised for the 4MB (or 16MB) of flash on the Waveshare module.

PicoMite software

The software loaded onto the RP2040-Plus includes the PicoMite firmware we introduced in July 2022 (Make it with Micromite). This is a powerful BASIC (MMBasic) interpreter for the Raspberry Pi Pico, with support for peripherals such as an LCD screen, SD card, sound etc. Because the BASIC language is built into the PicoMite, all games are written in BASIC.

The latest versions of the PicoMite firmware provide an A: drive, which acts like an SD card that cannot be removed. This allows us to store programs, music files, images etc internally, without the need for external storage like an SD card.

If you wish, you can plug an SD card (or microSD card in an SD card adaptor) into the socket on the LCD screen, and it will be available as “drive B:”. The menu system will

Games included in the Pico Gamer firmware



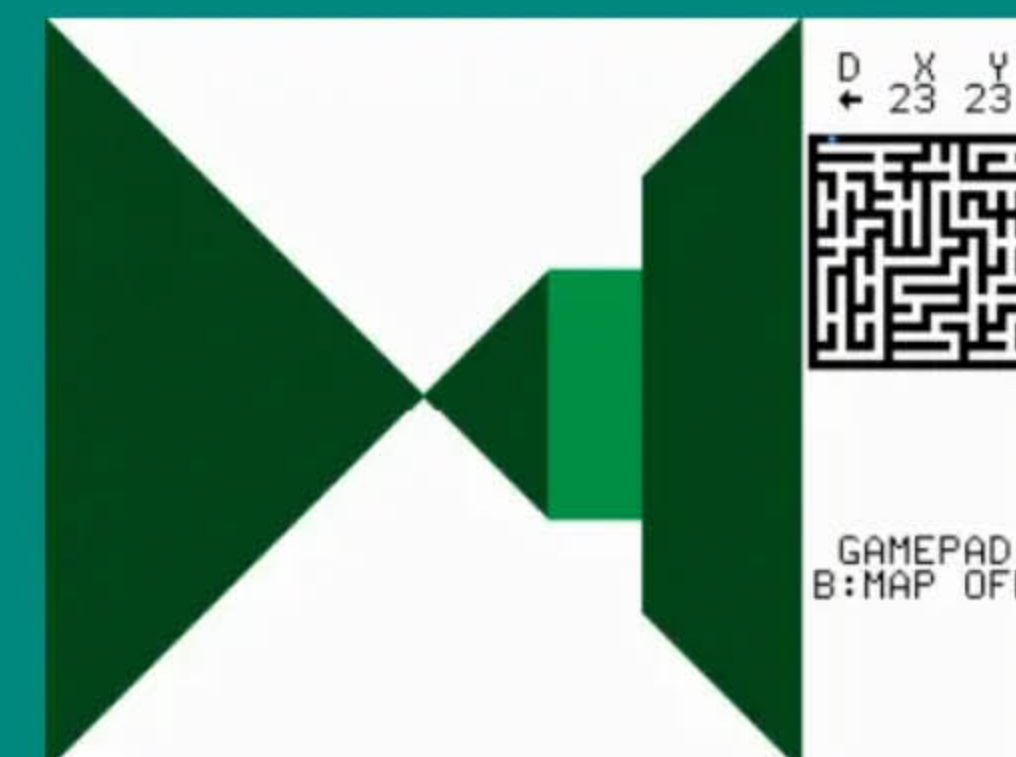
PETSCII Robots

In this complex strategy and exploration game, your goal is to enter the settlement and destroy the robots. The trick is finding the right tools and learning how to use them.



Lazer Cycle

You are on a fast Lazer Cycle, and so is your opponent. They will try to make you crash into a wall or track, and you try to do the same to them. It is a race to the death.



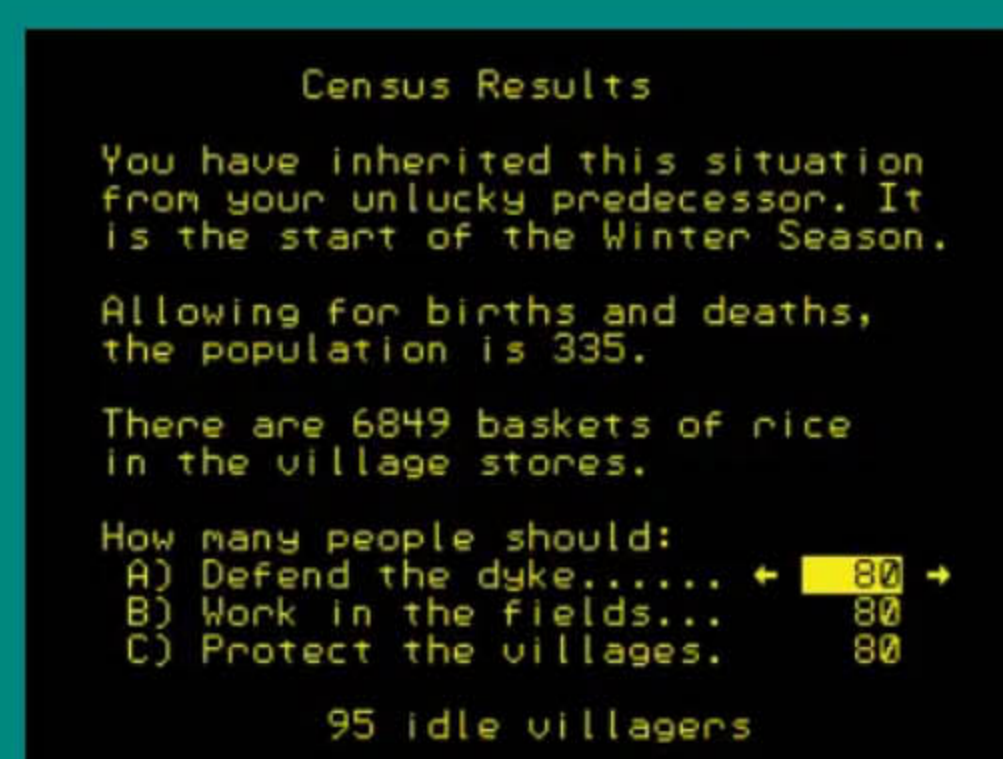
3D Maze

You are stuck in a 3D maze; your job is to explore and find a way out. You can call up a map to help, but it is not as easy as it looks.



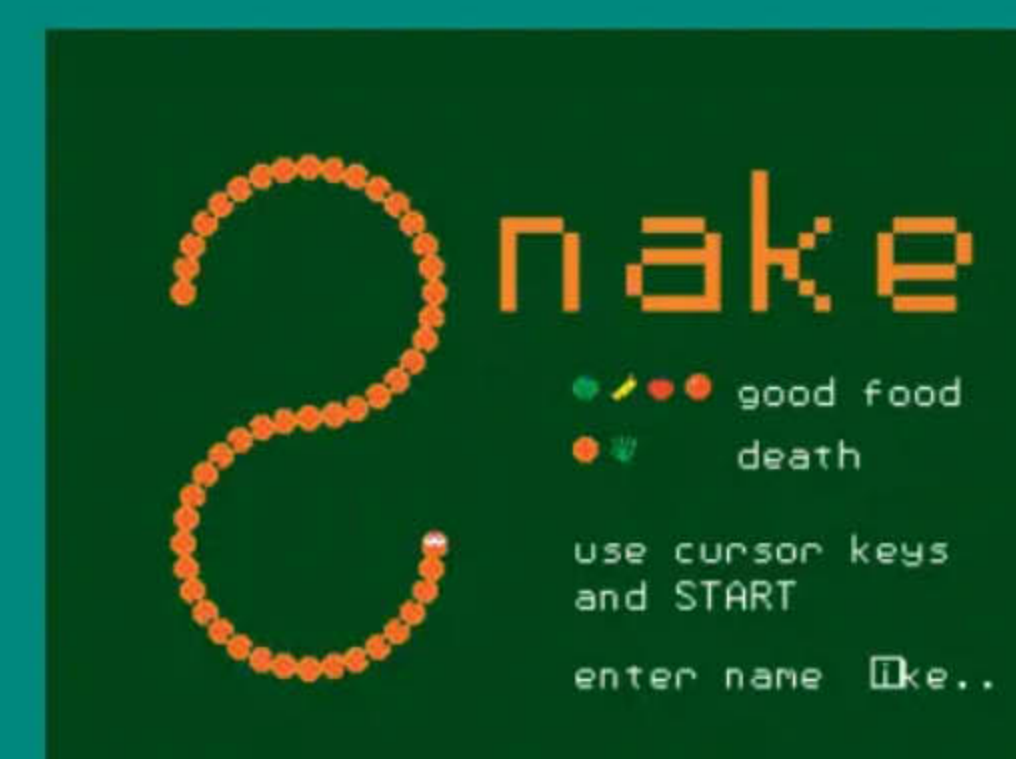
Pico Blocks

Similar to the ever-popular Tetris game, your job is to rotate and position colourful blocks falling from the sky into a neat carpet, where they will vanish. If your pile gets too high, you will lose.



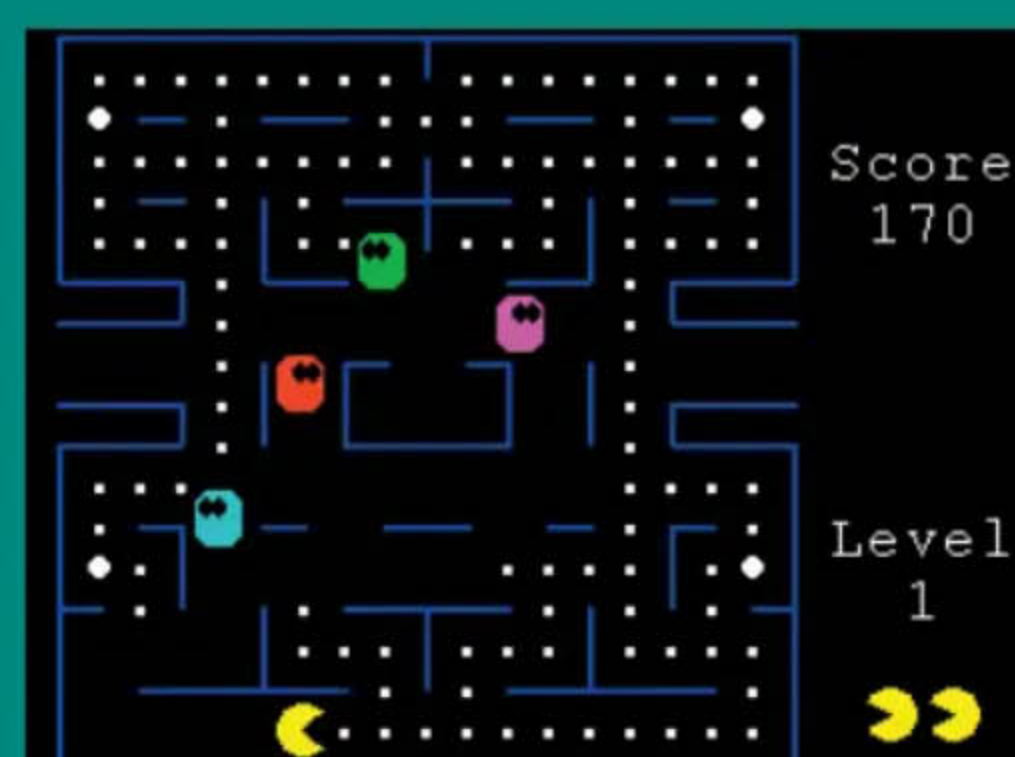
Kingdom

You are the ruler of the Yellow River kingdom and must allocate resources between feeding the hungry people & defending against thieves. Don't get it wrong, because the people might revolt.



Snake

Guide your snake around the board, eating the good food and avoiding the bad. As you eat, your snake will get longer. You will need all your skills to avoid crashing into a wall!



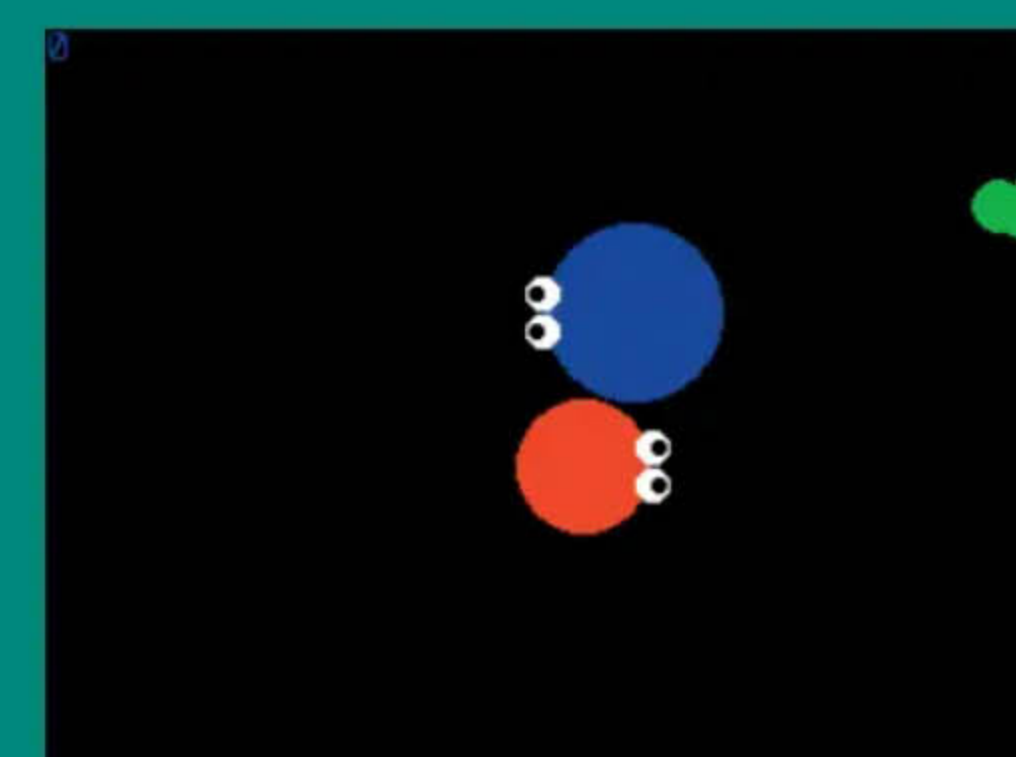
Pico Man

Inspired by the addictive Pac-Man game, you race around a maze, eating little dots while being chased by four ghosts. Eating a Power Ball gives you special powers, so you can pursue the ghosts instead!



Pico Vaders

Based on the classic Space Invaders from the 1980s, you are faced with hordes of invaders who drop a steady stream of bombs while you dodge back and forth with your cannon, trying to shoot them down.



Circle One

This game's objective is to eat the apples and grow while your opposition (the computer) will try to do the same. The one who grows to the maximum size wins. It is a simple but entertaining game that is perfect for young children.

allow you to switch to this drive and run games from it. However, this is not a requirement, and usually, the internal file system (drive A:) is sufficient to store all the games.

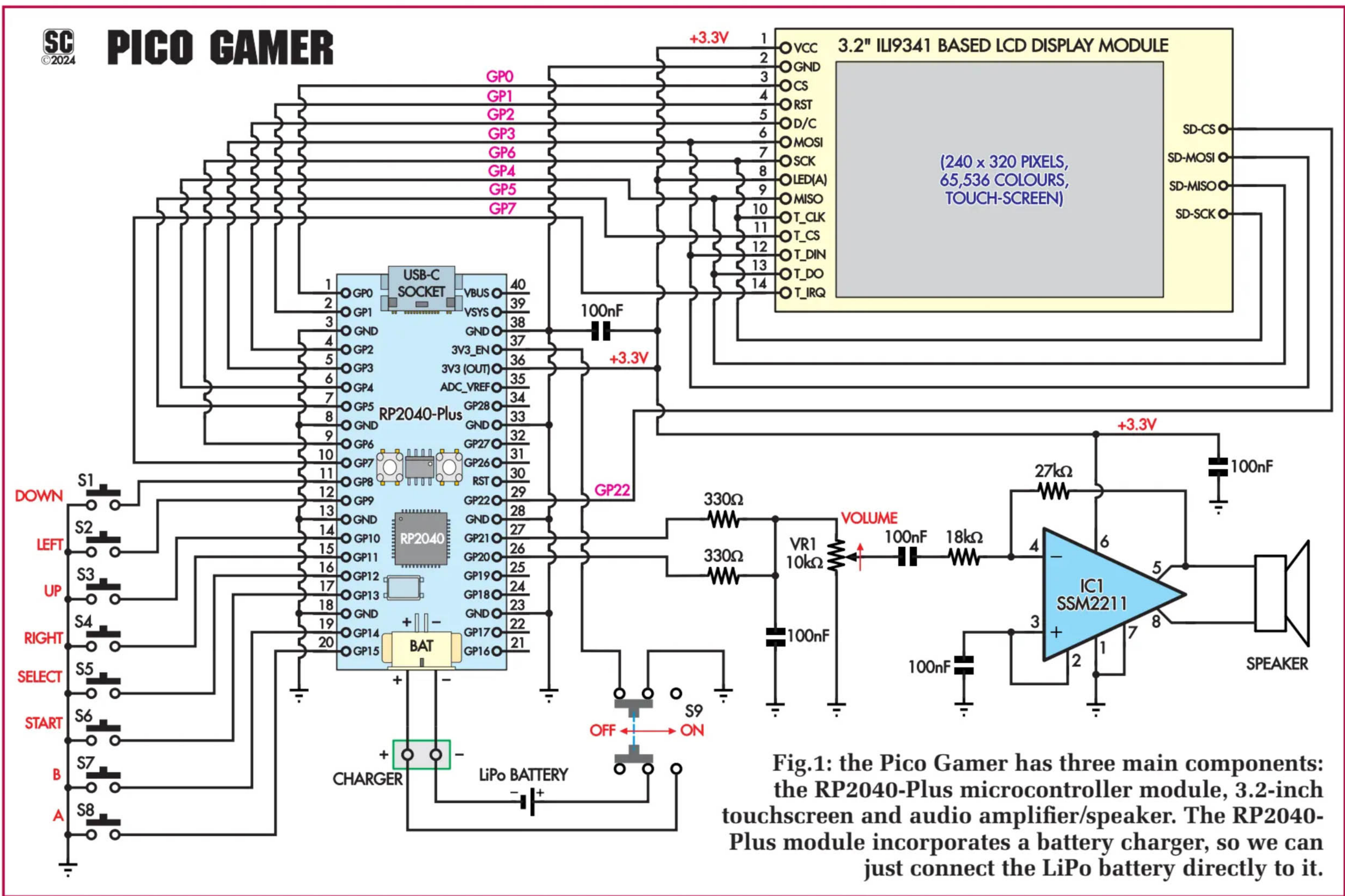
The PicoMite also implements flash slots as alternative storage places for programs. There are three of them, and when a program is run from one of these, it does not need to be loaded

into the main program memory, so it executes quickly. The Pico Gamer's menu program is stored in the first flash slot, so it is always ready to run.

Acknowledgments

The Pico Gamer is based on the work of many people from around the world. The concept of a Game Boy lookalike using the Raspberry Pi Pico

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was pioneered by Tom Williams in the UK, and he designed the Game*Mite with some help from Australian Mick Gulovsen.

He published his design on The Back Shed Forum (<https://pemag.au/link/absd>) and it has been quite successful, with several hardware clones and many extra games added to its repertoire.

The games themselves came from authors including Martin Herhaus (Germany), Harm (Netherlands), Tom Williams (UK) and Geoff Graham (Australia). Tom Williams also wrote most of the utility programs. You can contact these authors on The Back Shed Forum with suggestions and bug reports if you need to.

We have kept the hardware features of the Pico Gamer compatible with the Game*Mite, so games and programs written for one will run on the other. If you wish, you can even load Tom's full firmware package for his Game*Mite onto the Pico Gamer, and it will run equally well.

Circuit details

As you would expect, the circuit (Fig.1) is dominated by the RP2040-

Plus module. The eight game buttons connect directly to the processor, pulling the associated input pin low when pressed. Programs running on the Pico Gamer configure these pins as inputs with internal pullup resistors, so external resistors are not required.

The power switch in the off position disconnects the battery and disables the power supply in the RP2040-Plus. The latter is done so that the Pico Gamer will shut down even if it is connected to a USB power supply.

The battery charger in the RP2040-Plus will terminate at 4.2V, the correct voltage for standard LiPo batteries, so it will not overcharge them. When the Pico Gamer runs on battery power, the protection circuit within the battery will automatically disconnect the load so you cannot damage the battery by accidentally leaving the console on.

The LCD screen is connected to the processor via an SPI bus, which drives the display, touch controller and SD card socket. The LCD and the audio amplifier are both powered by the RP2040-Plus via its 3.3V output. This is used because the Pico's on-board DC-to-DC converter ensures a

constant output voltage regardless of the battery voltage, which can vary from 4.2V to about 2V.

The sound output is generated as stereo pulse width modulated (PWM) signals from digital output pins GP20 and GP21. These signals are filtered and summed by the two 330 Ω resistors and the 100nF capacitor. The resultant mono audio is fed to an SSM2211SZ audio amplifier, which drives the speaker in a bridged configuration.

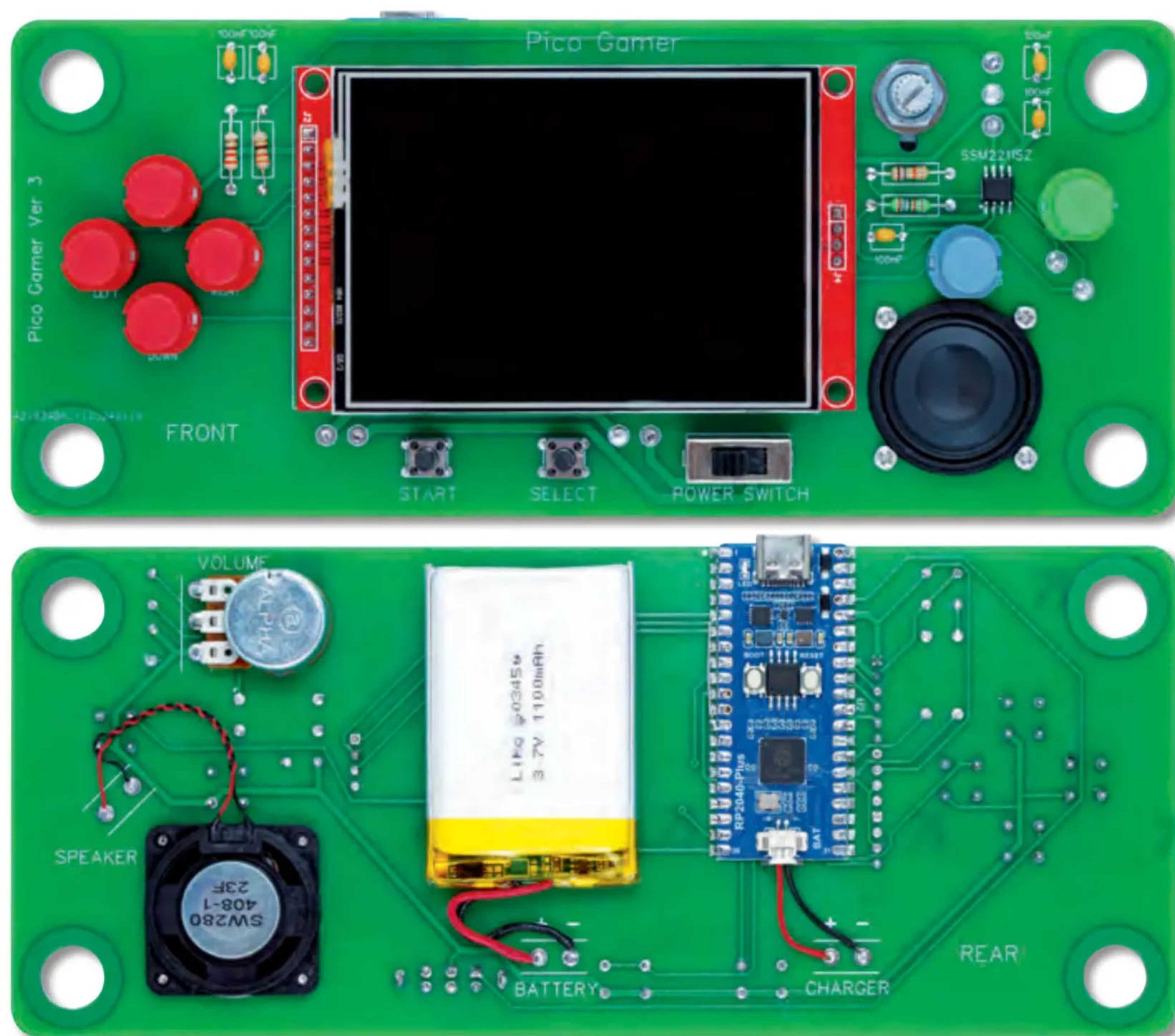
With a 3.3V power supply, the SSM-2211SZ does not generate much power, but the volume is ample for a handheld device.

Sourcing the parts

SILICON CHIP is offering kits that include all parts except the battery (which can't be sent by airmail). There is the option of no case (if you want to print your own), a basic case that you can paint any colour and a more expensive black case that shouldn't need to be painted. So that's one way to get the parts to build the Pico Gamer.

You can get a suitable 1100mAh 3.7V LiPo battery from a local seller on eBay.

The Pico Gamer



The front of the PCB has the 3.2-inch LCD screen, buttons, switches, audio amp IC & passive components. The rear of the PCB holds the RP2040-Plus, battery, volume potentiometer & speaker. It is necessary to solder the RP2040-Plus flush with the PCB so that the USB connector aligns with the cutout in the case.

If you want to gather the rest of the parts yourself, here are some suggestions:

The core of the Pico Gamer is the Waveshare RP2040-Plus, available from Waveshare (www.waveshare.com) and Amazon. You only need the 4MB version; make sure you purchase it without header pins, as it must be soldered flush with the PCB.

The battery charger socket on the RP2040-Plus is a two-pin Molex PicoBlade with a 1.25mm pitch. The matching plug with attached wires is commonly used in drones and can be purchased from drone suppliers. Note that many battery connectors on offer are JST-style connectors, such as JST-SH or JST-XH, which are incompatible.

Another way to get a matching connector is to buy a battery on eBay that comes fitted with a PicoBlade connector. You can then cut this off and use it as the charging cable, while the now unterminated battery leads can be soldered directly to the PCB.

The LCD is a 3.2-inch panel with a 320×240 pixel resolution using the ILI9341 controller. There are many on offer on eBay and AliExpress, but

make sure the vendor's photo matches Fig.4 (shown at the end of the article), as there are some incompatible designs that will not physically fit.

You can purchase the display without the touch interface, which would work fine as no games currently use that feature. However, you will only save about a dollar, so you might as well get it regardless.

The large, coloured tactile switches have 8mm diameter buttons and can be purchased from various international suppliers. We found one version from an Australian supplier that had a better 'clicky' feel, so that is what we have included in the SILICON CHIP kits, but your preference might differ.

The tactile switches for the start and select functions need a relatively long shaft of around 9mm, with a total height of 13mm (including the button base). These can be found on eBay and AliExpress.

The volume potentiometer is a standard 16mm logarithmic type and is available on eBay and AliExpress. The value is not critical; it can be in the range of 10kΩ to 50kΩ, but its depth must be less than 10mm to fit

in the case, and it should have an 8mm-long knurled shaft as it is used without a knob in this design.

The loudspeaker used in this design is the DB Unlimited SW280408-1 (Mouser Cat [497-SW280408-1](#), DigiKey Cat [2104-SW280408-1-ND](#)). This was chosen as it's small but has decent sound quality and is easy to mount using four small screws.

Even if you don't have a 3D printer, getting custom-designed 3D-printed case pieces is relatively easy. The two STL files defining the top and bottom halves of the case can be downloaded as part of this month's package on the Electron Publishing website and sent off for fabrication.

There are numerous online 3D printing services but we recommend JLCPCB. You only need to upload the files to their website and select their SLA process using LEDO 6060 resin (<https://jlc3dp.com/3d-printing-quote>). They will then make and ship the case to you within a few days.

The 6060 resin is strong, with no warping, and the surface is smooth in a slightly translucent off-white colour. However, note that this material can yellow slightly with age, so you might want to spray paint it. The 6060 resin readily accepts paint. An ideal paint for this purpose is Rust-Oleum Satin 2X Ultra Cover, available from large hardware stores in many colours.

Alternatively, you could use one of JLCPCB's more expensive materials that are dyed or otherwise immune to yellowing, for example, "Black Resin" or "Imagine Black". We offer one of those options in our kits for those who don't want to mess around with paint and like the 'stealthy' appearance.

Construction

Only a few components are involved in the Pico Gamer, so construction can be completed in an hour or two. Four components (the RP2040-Plus, battery, volume potentiometer and speaker) mount on the rear of the PCB, with the rest on the front side. The PCB is marked FRONT and BACK to help with the orientation.

The Pico Gamer PCB is coded 08104241 and measures 188 × 80mm. During construction, refer to the overlay diagrams, Figs.2 & 3, to see which parts go where. You can also check the photos.

Constructional Project

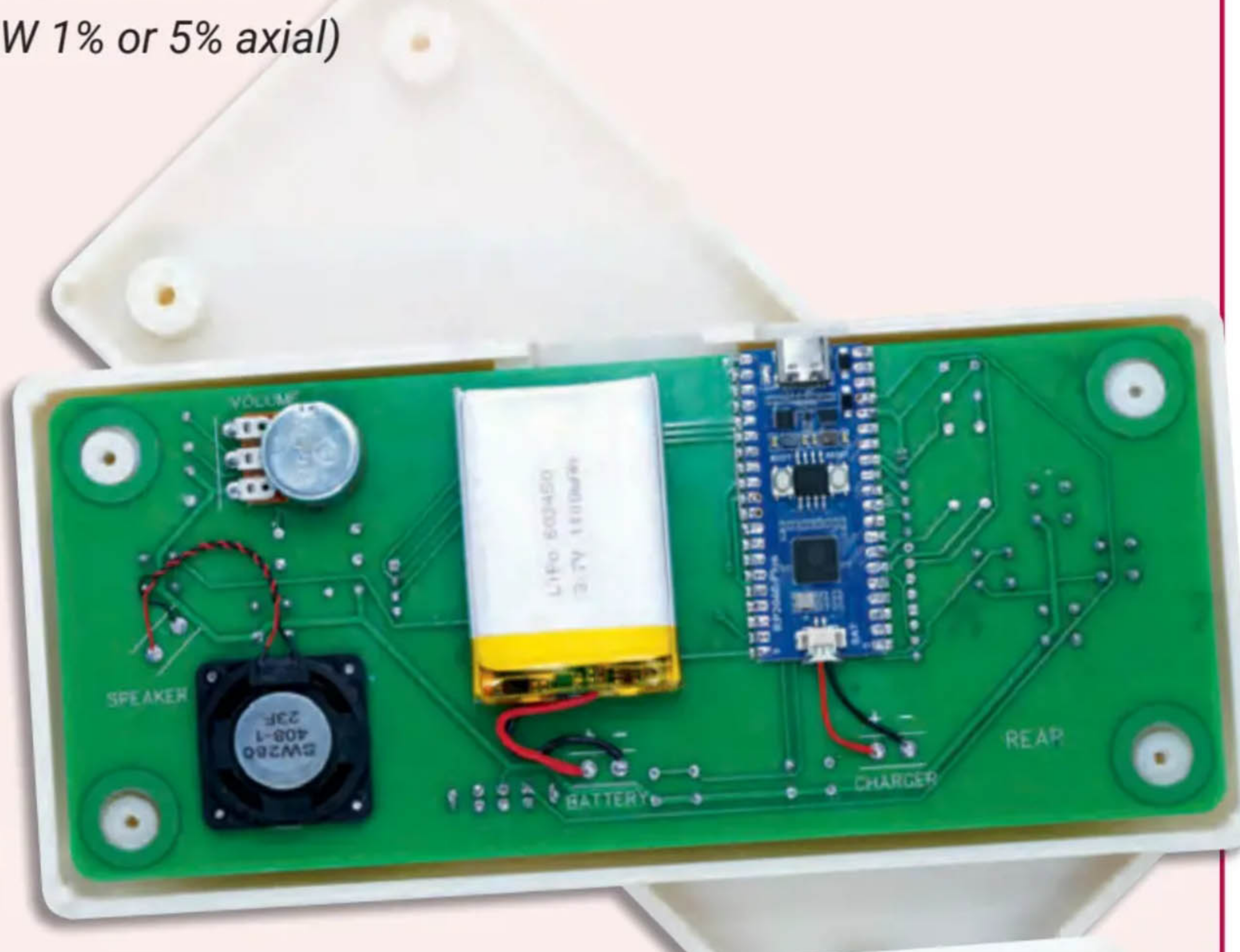
Parts List – Pico Gamer

- 1 double-sided PCB coded 08104241, 188 × 80mm
- 1 custom 3D-printed case in two pieces (upper and lower), 199×90×26mm (see text)
- 1 Waveshare RP2040-Plus module with 4MB or 16MB flash memory, without header pins [**Waveshare SKU 20290 (4MB) or 23503 (16MB)**]
- 1 3.2in LCD touchscreen, 320×240 pixels, with ILI9341 controller and SD card socket
- 1 900-1100mAh 3.7V LiPo cell
- 1 SSM2211SZ 1.5W amplifier, SOIC-8 (IC1) [**DigiKey, Mouser, RS, Farnell**]
- 6 SPST momentary tactile switches with 8mm diameter buttons, 5×5mm pitch, in various colours (S1-S4, S7, S8) [eg, C&K D6 series @ **Farnell**]
- 2 SPST momentary tactile switches, 4×6mm pitch, 13mm height (S5, S6)
- 1 PCB-mount miniature DPDT slide switch (S9)
- 1 DB Unlimited SW280408-1 8Ω loudspeaker [**Mouser 497-SW280408-1, DigiKey 2104-SW280408-1-ND**]
- 1 10kΩ logarithmic potentiometer with 8mm spline shaft
- 5 100nF 50V X7R multi-layer ('monolithic') ceramic capacitors, 5mm pitch
- 1 4-pin header, 2.54mm pitch
- 1 2-pin Molex PicoBlade plug, 1.25mm pitch, with attached leads
- 4 M3 × 16mm panhead machine screws
- 4 M2 × 6mm panhead machine screws
- 1 can of spray paint (**optional**; see text for recommendations)
- 1 double-sided foam adhesive tape strip or pad

Resistors (all 1/4W 1% or 5% axial)

- 1 27kΩ
- 1 18kΩ
- 2 330Ω

The PCB fits neatly into the 3D-printed case. When the two halves of the case are screwed together, it has the optimal dimensions for a handheld game console with a smoothly rounded shape that fits well in the hands.



With its custom 3D-printed case, the Pico Gamer is a professional-looking game console. It comes with nine games, including some inspired by Pac-Man, Space Invaders and Tetris, that work well with its colourful 3.2-inch LCD screen. The inbuilt rechargeable battery lets you play for up to eight hours at a time.

Start with the SSM2211SZ audio amplifier chip, which is in a small 8-pin surface mount package that is much easier to fit when no other components are in the way. This mounts on the front side of the PCB and should be soldered using the standard technique for SMD ICs.

Apply a little flux paste to the PCB pads and place a small amount of solder on a corner pad. Position and hold down the IC, observing the dot marking pin 1, and tack solder one of the pins using the solder on the pad. Check and correct the IC's alignment, then tack solder the pin in the opposite corner.

With the IC secured, apply more flux paste and, with the bare minimum of solder on your iron, place its tip on the end of each pin, letting the solder flow around the pin and the solder pad. Finally, inspect your work with a strong magnifier (×10 or ×20) and correct any problems with more flux paste and solder-wicking braid.

Next, you should install the RP2040-Plus on the rear side of the PCB. This sits flush on the PCB, making it a surface-mounted component. Ensure that it is aligned centrally on the solder pads and that the USB socket is at the top, protruding over the edge of the PCB.

The battery charger plug and cable can be soldered now. Note that the colour of the wires (red/black) crimped to the connector might not match the polarity marked on the RP2040-Plus. Check this, and make sure that the lead from the + side of the connector goes to the pad marked + on the PCB regardless of the wire's colour.

Next, fit the resistors and capacitors. There are nine in total, and none are polarised, so installation should be easy. The parts list includes resistor colour codes, but you can also use a DMM set to measure ohms to verify their values.

Installing the LCD screen

The next component to install should be the LCD panel. For height reasons, it is not socketed; instead, the pin headers go through the holes in the PCB and are soldered on the other side. These displays are notoriously sensitive to static discharge, so make sure that you ground yourself before unwrapping it and avoid handling it too much, especially its connecting pins.

The Pico Gamer

Most LCD panels are supplied with the main connector header pins installed, but you will need to add a four-pin header for the SD card interface in the locations marked SD-CS etc. Then insert the LCD panel into position on the front side of the PCB and push it down until it is flush with the PCB.

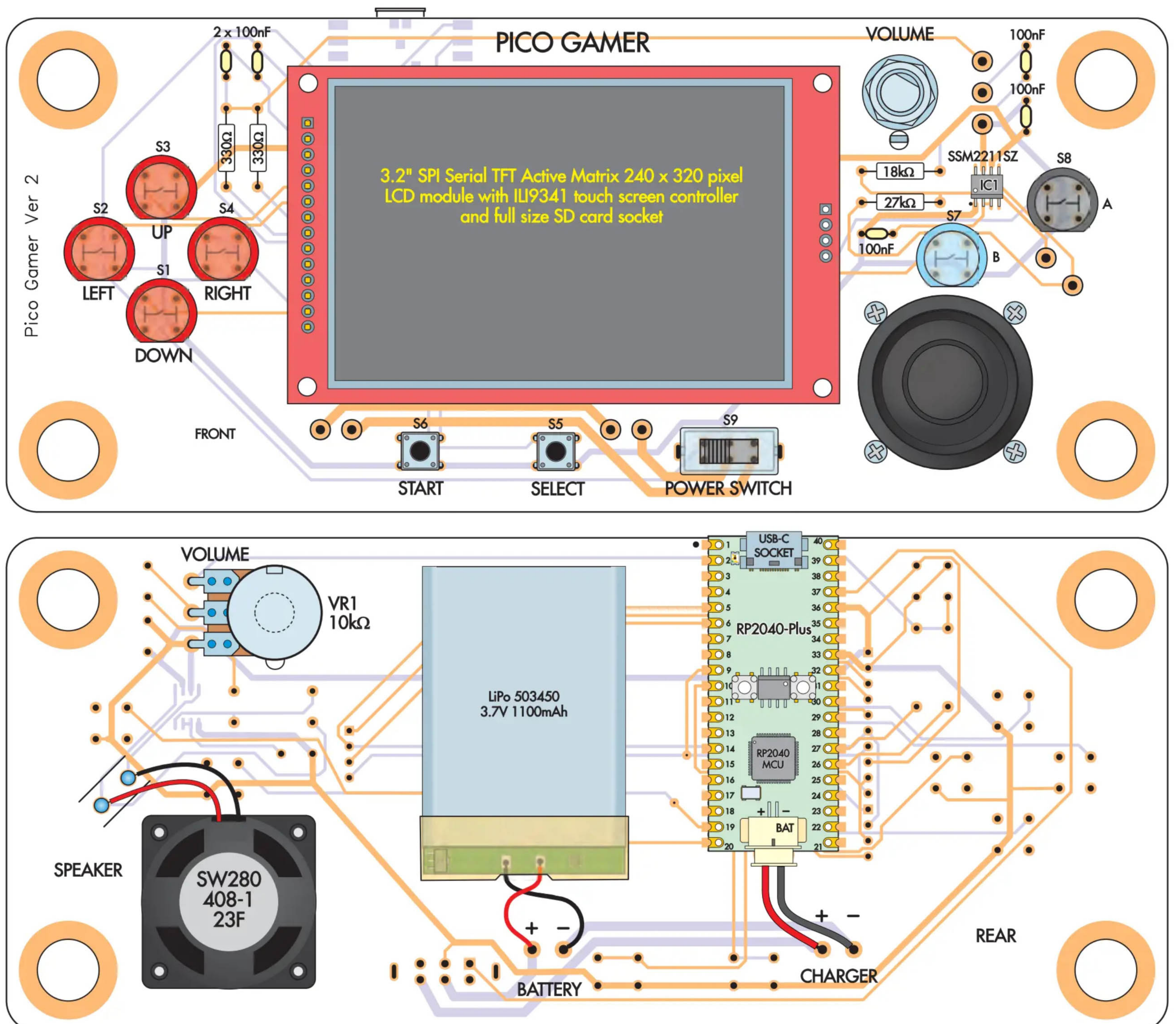
Turn the PCB over and temporarily place it in the top section of the 3D-printed case, ensuring it sits correctly on the four mounting pillars. Next, push down on the LCD screen's header pins until the LCD glass is flush with the case's front bezel. You can then solder and trim the pins.

The reason for this operation is to ensure that the LCD's glass will sit flush with the front bezel while compensating for minor variations in the 3D printing of the case. Don't force the LCD hard against the case, as that could interfere with the touch function; a flush contact is all that is required.

Next, fit the volume potentiometer by inserting it through the back of the board, with its locating pin in the hole provided, then tighten the supplied nut over the washer to hold it in place. After that, bend the solder tabs towards the PCB and directly solder them to the pads provided.

Now attach the speaker to the rear of the PCB, with the front of the speaker cone facing through the hole. The speaker is held in place by four M2 machine screws inserted from the front side of the PCB which self-tap into the speaker's mounting holes; nuts and washers are not required. Once it is securely in place, solder its two wires to the nearby terminals.

After that, mount the coloured button switches, the long shaft tactile switches and the power switch, all on the front of the PCB. You can use whatever coloured buttons you want; our kits will come with the same set shown in the photo below. If you order the kit with



Figs.2 & 3: there are components mounted on both sides of the board. As there are not too many, it shouldn't take long to assemble. They are all pretty easy; IC1 is surface mounting, but its pins are wide enough to be soldered individually.

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the dark grey/black case, the black button will be white instead for better contrast.

The battery is soldered last. Secure it in its marked position on the rear of the PCB using double-sided adhesive foam tape. Kits will include a rectangular pad that might need to be cut down if it's too large.

The battery (cell) will typically come with a connector that needs to be cut off and the leads directly soldered to the PCB. Remember that the battery will be supplied partially charged, so do not accidentally short its leads when trimming them (cut them separately) and ensure that the power switch is in its off position before soldering the battery to the PCB.

Also ensure that one lead doesn't accidentally come in contact with the opposite lead or pad while you are soldering.

Before installing the PCB in the case, if you have the off-white (6060 resin) case, both halves can be spray-painted in your choice of colour. This allows you to customise the case and protects the plastic from the environment so it won't go yellow over time.

The two halves of the case are held together by four 16mm-long M3 machine screws inserted in the bottom half of the case that self-tap into the plastic of the top case. This works well,



On the rear of the Pico Gamer is the USB Type-C connector for charging or connecting to a desktop computer for firmware updates and writing games. There is also a full-size SD card slot for extra game storage (microSD cards can be used with commonly available adaptors).

but it is not strong enough to survive repeated disassembly, so make sure that you test the completed PCB before you screw the case together.

Loading the firmware

Installing the firmware on the Pico Gamer is easy. All you need to do is press the left-hand button (marked Boot) on the RP2040-Plus and plug the USB interface into your desktop or laptop computer while holding down that button.

The RP2040-Plus will connect to your computer and imitate a USB flash memory drive. Most computers will then open a file management window showing the contents of this drive (which you can ignore).

The Pico Gamer firmware file can be downloaded from <https://pemag.au/Shop/6/370> or <https://geoffg.net/picogamer.html>

It will have a filename similar to "PicoGamerV1-2.uf2". There will be two versions, one for 4MB modules and one for 16MB, so select the appropriate one for your build. Drag and drop that file into the imitation USB drive created by the RP2040-Plus. When the copy has finished, your Pico Gamer will reboot and display the main menu.

The firmware file you uploaded contains everything needed, including the BASIC interpreter with all the necessary options set, the menu program stored in flash slot 1 and the internal file system with all the games and their supporting files (images and music). There is nothing extra that you need to install or configure. Just start playing.

In the future, you can update the firmware without opening the case by connecting the USB socket to a desktop or laptop computer and, using a terminal emulator, interrupt the running program by pressing CTRL-C, then enter the following command:

UPDATE FIRMWARE

Running this command will have the same effect as disassembling the case and holding the Boot button on the RP2040-Plus.



The original colour of the case is white as shown in this photo. The case in the lead photo was spray painted in the "Satin Claret Wine" colour. For the buttons, you can choose whatever colours take your fancy. This main menu is displayed immediately after powering on the Pico Gamer. The up/down buttons allow you to select a game and pressing SELECT will run it. If you choose a directory, the contents of that will be displayed instead. When an SD card is inserted, you will have the option of playing games from it too.

The Pico Gamer

Using the Pico Gamer

The first thing that you need to do is charge the battery. To do this, connect the USB socket on the rear of the console to a charger via a USB-C cable and switch the Pico Gamer on. The charging time from completely flat is about four hours.

When the Pico Gamer is running on battery and the menu is displayed, an estimate of the battery's charge (high, medium or low) will be displayed on the bottom line of the main menu.

When you turn the Pico Gamer off, the battery is completely disconnected so that it will keep its charge for a long time if it is not used (a year or two). Because the battery is disconnected when the device is off, it will only charge when the device is switched on and plugged into power.

Also on the rear of the case is the slot for a full-size SD card. Cards up to 32GB formatted in FAT16 or FAT32 can be used, and when a card is inserted, the main menu will show an option for selecting it (button B).

When the SD card is selected, the menu system will show the directories and executable files on it in the same format as the internal file system. The A button allows you to swap back to the internal file system if needed.

Whenever the Pico Gamer is powered up, it will run the menu program (in flash slot 1) and display the main menu. Using the menu is intuitive – you use the up/down buttons to select a game or program and press select (SEL) to run it. Subdirectories are also shown in the menu list, and if you choose one of them, it will show the contents of that directory.

In most games and within the menu, the select (SEL) button on the front (under the LCD screen) is used to select an option or exit the current mode. The START button is generally used to start a game running or pause it if running, although that can differ between games.

The functions of the other buttons are defined by the game. Typically, the four navigation buttons (on the left) are used to move in various directions, while the A and B buttons (on the right) fire guns, drop bombs or perform other similar functions.

The easiest way to permanently install a new game is to copy it to an SD card and insert the card into the Pico Gamer. Then plug the USB

The 3D-printed case

While a few of our projects have used a custom 3D-printed case before, this is the largest and most complicated one so far. The good news is that, besides being able to get the case in kits, companies also exist that can do the printing for you. And because they use large industrial machines and a wide range of materials, they can do a great job.

In the past, we would make our own PCBs at home, including the cutting, etching and drilling. Few people would do that these days because commercial companies do such a good job fabricating PCBs at a very competitive price. 3D printing is starting to go the same way.

A custom case has many advantages over buying a standard ABS plastic box. For a start, it has the optimal dimensions for the design, and it can have a smoothly rounded, ergonomic shape.

It can also have the correct holes and cutouts precisely positioned, and they will be perfectly formed – not the jagged mess that can happen when they are cut by hand.

Another benefit is the thin and professional bezel around the LCD panel. That has long bedevilled designs incorporating an LCD panel – achieving the perfect thin rectangular bezel by hand is tough. However, it can be done easily with a 3D-printed case.

A typical medium-sized 3D-printed case will cost \$20-50 to make, which is good value given that you get precisely what you want with all the holes and various features properly made for you.

Design software

If you wish to design your own case, you first need to decide on the 3D modelling software you will use. There are quite a few free packages to choose from. However, many are not quite as intuitive and accomplished as modern PCB design packages.

Suitable packages range from Tinkercad, which runs in a browser and is aimed at beginner and educational users, through to more heavy-duty offerings such as Blender. Blender is free and open source; it is very capable and can do almost anything. It also has a very steep learning curve with many 'gotchas' and non-intuitive operations that can trip you up.

We settled on Blender because we needed to create a very smooth bevel on the corners of the case, and we wanted to position features with a precision of a fraction of a millimetre. However, the steep learning curve caused us to pull out a lot of our hair in the process.

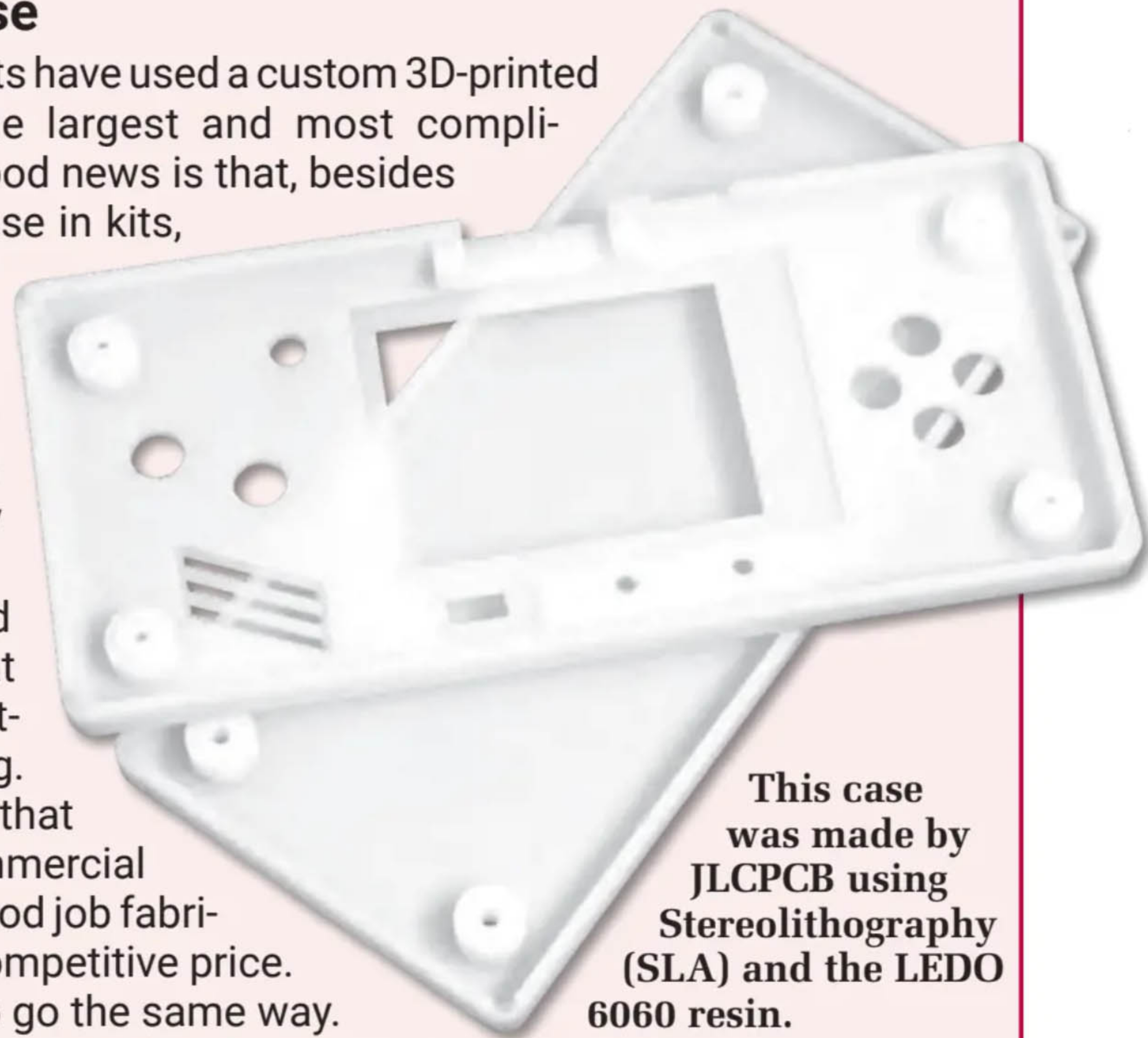
If you plan to use Blender to design an instrument case, we recommend this tutorial video as it covers most of what you need to know: <https://youtu.be/rN-HMVTB7nk>

Fabricating the case

When you have finished the design, you can export it as an STL file and send it to your fabricator of choice. There are many, both within Australia and overseas.

We used JLCPCB in China, who are better known for producing PCBs but now they are moving into 3D printing. They offer quite a few technologies and materials, including 3D printing in metal.

We chose their stereolithography (SLA) technology for the prototype, using the LEDO 6060 resin. This is one of the more cost-effective processes, and it produced a perfect result with all the special features positioned with pin-point accuracy.



This case was made by JLCPCB using Stereolithography (SLA) and the LEDO 6060 resin.

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into a desktop or laptop and, using a terminal emulator, interrupt the running program on the Pico Gamer with CTRL-C (full instructions are in the PicoMite User Manual).

This will return you to the command prompt, where you can copy the game to the Pico Gamer's internal file system with the command:

COPY "B:filename" TO "A:"

To return to the menu, enter:

FLASH RUN 1

You will see your new game included in the menu.

Writing games

One of the best features of the Pico Gamer is that you can write your own games for it. It is not hard to do and can be a great learning experience.

Programs are written in the BASIC language, which is easy to learn by design. The version of BASIC running on the Pico Gamer (MMBasic) has many features for displaying graphics and text, playing various sounds (including music) etc. To learn about this, download the PicoMite User Manual from the bottom of <https://geoffg.net/picomite.html>

There is an active community on The Back Shed Forum for people who are writing games for the Game*Mite and the Pico Gamer and posting them on the forum (www.thebackshed.com/forum/Microcontrollers). If you write a game, you should join the

forum and post your efforts there for others to enjoy.

The best way to write your games is to plug the Pico Gamer's USB port into your desktop or laptop computer. You can then use a terminal emulator program such as Tera Term to gain access to the PicoMite's console and use the editor built into MMBasic to enter the program.

The PicoMite User Manual goes into more detail on how to use the editor; it is a very efficient process with the ability to quickly jump between running the program and editing it.

Another method is to use MMEdit, a program that runs on your PC and allows you to edit the program on the PC, then transfer it over USB to the Pico Gamer and run it with a single keypress. MMEdit is written by Jim Hiley in Tasmania and you can download it from www.c-com.com.au/MEdit.htm

Detecting button presses

When you write a game for the Pico Gamer, you need to keep a couple of things in mind. The first is how to detect a button press.

The eight buttons on the Pico Gamer are connected to I/O pins GP8 to GP15 (physical pins 11 to 20) and will pull the pin low (ie, logic zero) when pressed. The first thing that your program needs to do is configure these pins as digital inputs with internal pullup resistors. For example:

```
For i = 11 To 20
  On Error Skip
  SetPin i, Din, PullUp
Next i
```

The "On Error Skip" command is necessary because some pin numbers in the range are ground pins, and MMBasic will throw an error when we try to configure them (or you could change the code to avoid the SetPin command for i=13 and i=18).

To detect which button has been pressed, you can define a series of constants representing the pin numbers like this:

```
Const bDOWN = 11
Const bLEFT = 12
Const bUP = 14
Const bRIGHT = 15
Const bSELECT = 16
Const bSTART = 17
Const bB = 19
Const bA = 20
```

Then, in your program, it is easy to determine if a button is pressed (pulled to a low level). For example:

```
If Pin(bDOWN) = 0 Then
  'down button pressed
If Pin(bLEFT) = 0 Then
  'left button pressed
```

Exiting to the main menu

The second feature that all programs must implement is to return control to the main menu when the user has finished playing and wants to exit. To do this, you need to insert the following command in your program:

Flash Run 1

MMBasic will immediately transfer control back to the menu program in slot 1 when this command is encountered.

The final step is to install your program on the Pico Gamer. To do this, you simply copy it as a file to the directory called /GameMite in drive A:, ie, the internal file. You can use the XModem command or copy the file to an SD card and use that to transfer the file to the Pico Gamer. When the main menu program runs, it will scan drive A:, discover the new program and include it in the menu.

So there you go, a modern handheld game console that encourages you to write your own games. For software updates, keep an eye on: <http://geoffg.net/picogamer.html>

PE

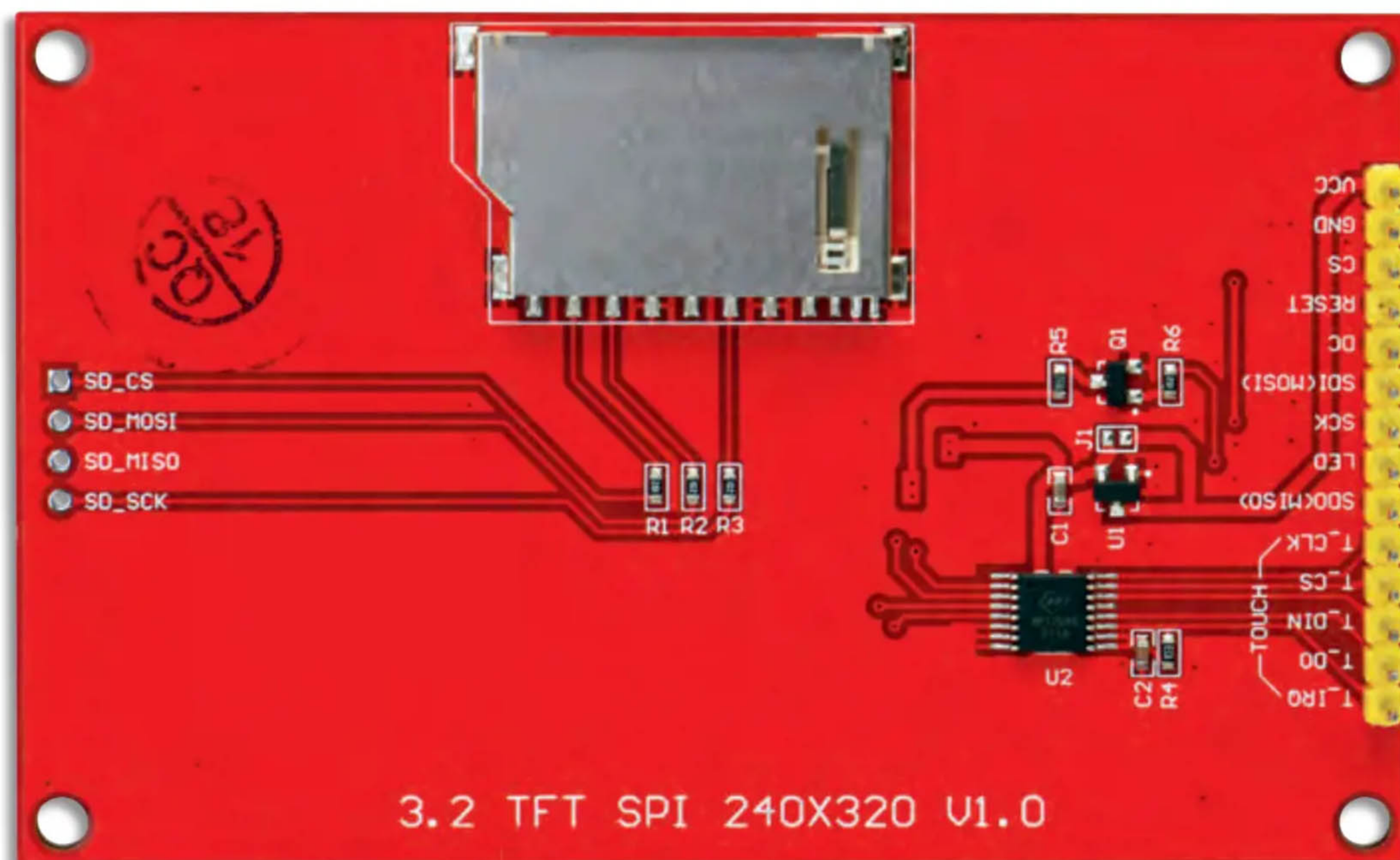
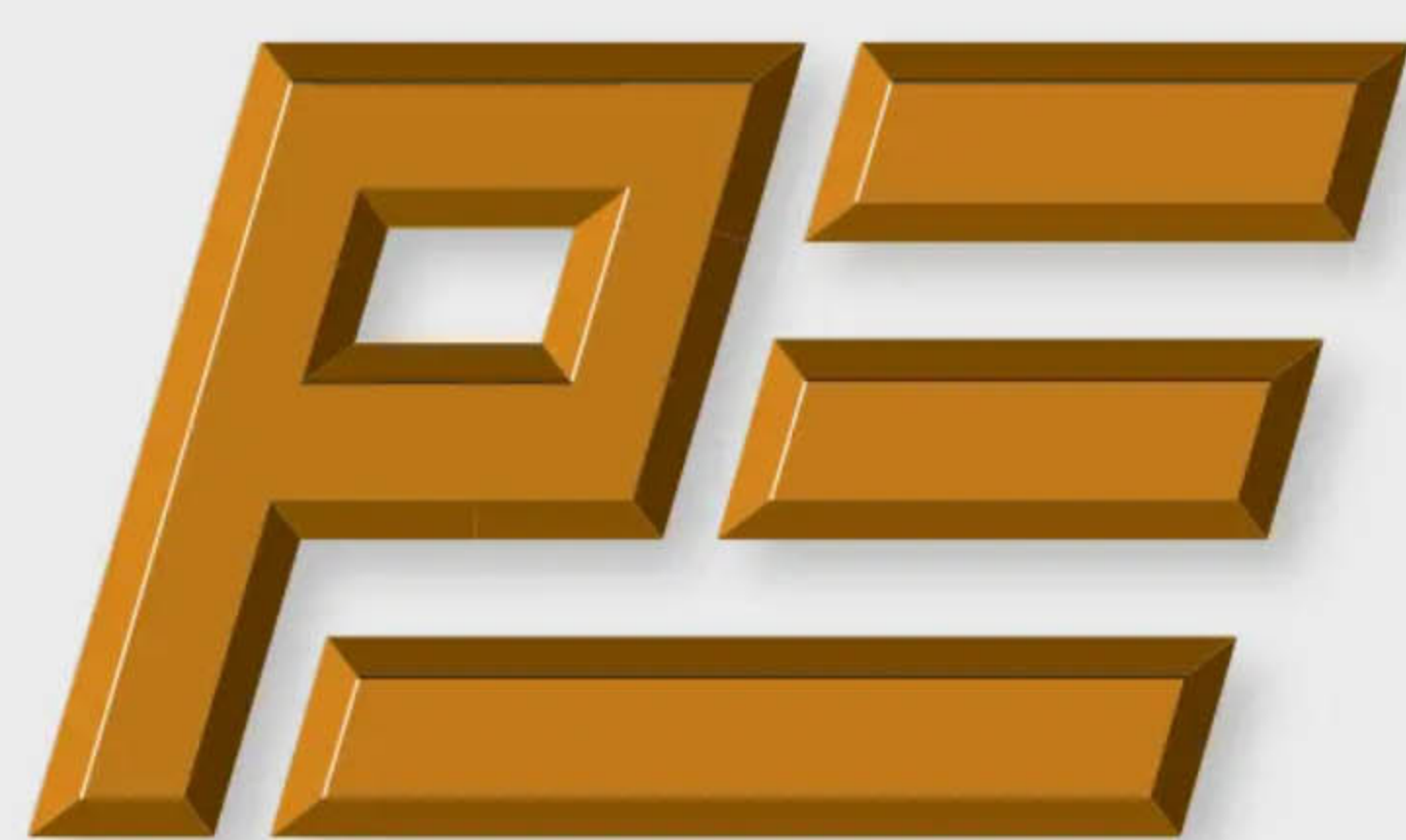


Fig.4: this is the LCD you should purchase for the Pico Gamer. They are available on eBay and AliExpress, but check that the vendor's photo matches this image, as some incompatible designs on offer will not physically fit. The front of our display reads "HR4 8637S G6/2" along the touchscreen.



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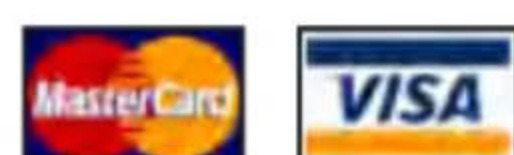
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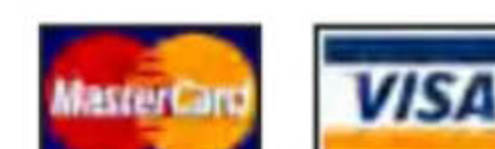
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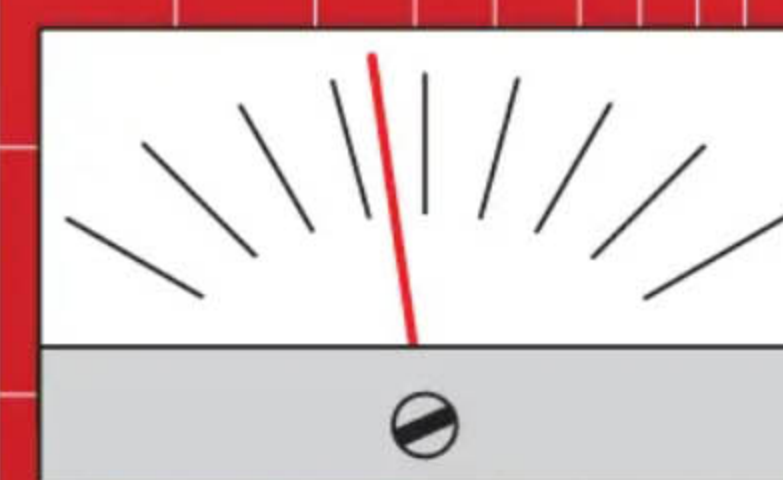
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AUDIO OUT



By Jake Rothman

Using Capacitors for HiFi, Part 1

Having read the editor's article "All About Capacitors" in the December 2024 issue, I thought a bit more detail could be added about their audio aspects, which differ from other areas of electronics. That took me down quite a rabbit-hole, as you shall discover!

Because high-quality audio is a relatively small area of the industry, audio engineers have had to develop their own specific capacitor knowledge base. This is almost passed down in the form of folklore. Consequently, there is a huge amount of "audiofoolery", rip-offs and falsehoods out there.

It's a bad idea to ask about capacitors on audio forums. They tend to be populated by musicians and DIY HiFi buffs who will put subtle artefacts and their own biases before rational engineering. It's quite likely you will find there somebody who suggests recapping your whole system with mil-spec Teflon capacitors.

Quotes such as "I had my whole mixer recapped by Quantum Dielectric Energy Systems.com and it sounds wonderful and it only cost me \$2000" are common. Often, so-called upgraded PCBs have mechanical problems. A classic error is hanging an 80mm-long axial

polypropylene capacitor on the pads designed for a 2.5mm-pitch electrolytic.

However, within this mass of falsehoods are some important truths and circuit techniques. Capacitor sonics are a subtle second-order effect. Capacitor distortion was once a contentious issue; traditional objective engineers like Peter Baxandall and Douglas Self didn't think it was significant enough to quantify.

It was only when Douglas Self, with his Blameless amplifier, got the total harmonic distortion (THD) level down to 0.001% that the capacitor-induced tip-ups in distortion at the low and high frequency ends were noticed and investigated

Engineers such as Ben Duncan and Graham Nalty in the 1980s (who've written for PE) also noticed capacitor effects, but the necessary instrumentation, such as the Radford Low Distortion Oscillator (LDO), wasn't readily available.

Capacitor non-linearity

The main concern for audio engineers is avoiding non-linearity, which manifests itself as harmonic distortion. This does occur with capacitors, but they are very low down on the list of

contributors. Loudspeakers are the worst offenders, with semiconductors and circuit design next.

Once there are enough active devices with plenty of negative feedback (NFB) in a good design, the total harmonic distortion due to these can drop below 0.001% (sometimes approaching even 0.0001%, which is just 1ppm!). This is the point where capacitor distortion can easily dominate that part of the circuit.

This is because capacitors are usually not within the negative feedback loop, so are not linearised. Even if they are, because an amplifier's open-loop gain is finite, the effects of their non-linearities on signals can't always be fully cancelled.

The subjective effects of capacitors are also overstated, especially since the distortion produced tends to be benign second and third harmonics, as opposed to distortion from op-amps and other high NFB amplifiers, which generate a spray of dissonant higher order harmonics.

I get annoyed when I test an expensive piece of audio equipment exhibiting distortion in the region of, say, 0.05% or more and finding expensive high-linearity capacitors fitted. What's the point, besides marketing? They would have been better off spending that money on something else!

Dielectric absorption (DA)

Known as DA or 'soakage', this effect is a slow release of charge as dielectric molecules return to their original random alignment after initially discharging a capacitor. This is why a capacitor that has been discharged can charge itself up a bit again later. In audio work, this can cause a small amount of very low-frequency shift that is subjectively unnoticeable, in my opinion.

In electrolytic capacitors, the DA can be as much as 3%, whereas for plastic-film types, it is much lower. It can be hidden in electrolytics due to leakage. The rectification, storage and

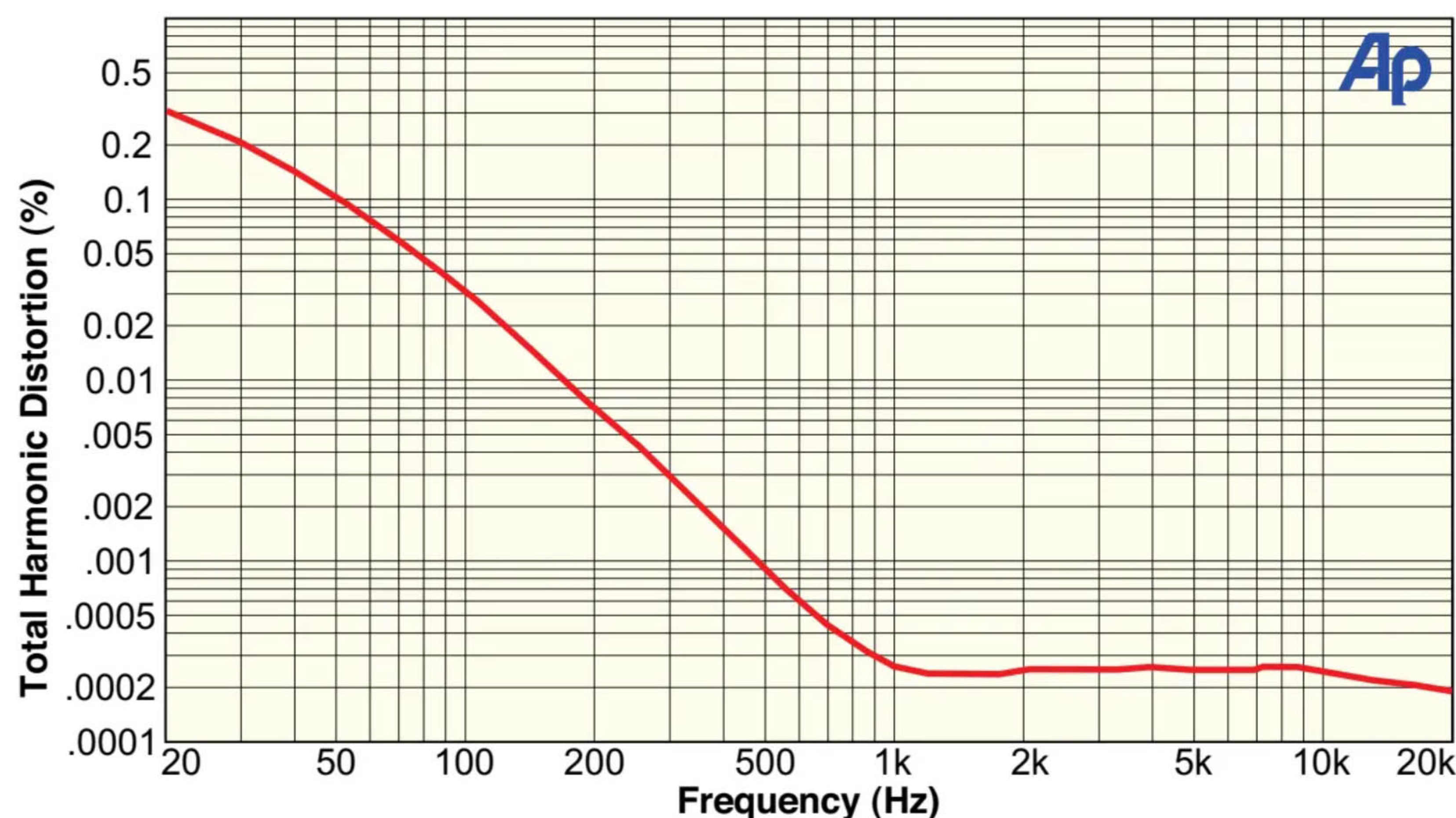


Fig.1: bad capacitor distortion: the typical rapid rise below 1kHz from a 4.7 μ F 10V cheap tantalum bead. This was 1V RMS into 600 Ω load; the -3dB point is 56Hz.



Fig.2: a component linearity tester at Charcroft's capacitor and resistor plant in Wales. It screens for bad contacts, which cause distortion.

cone-drift effects in loudspeakers are much worse.

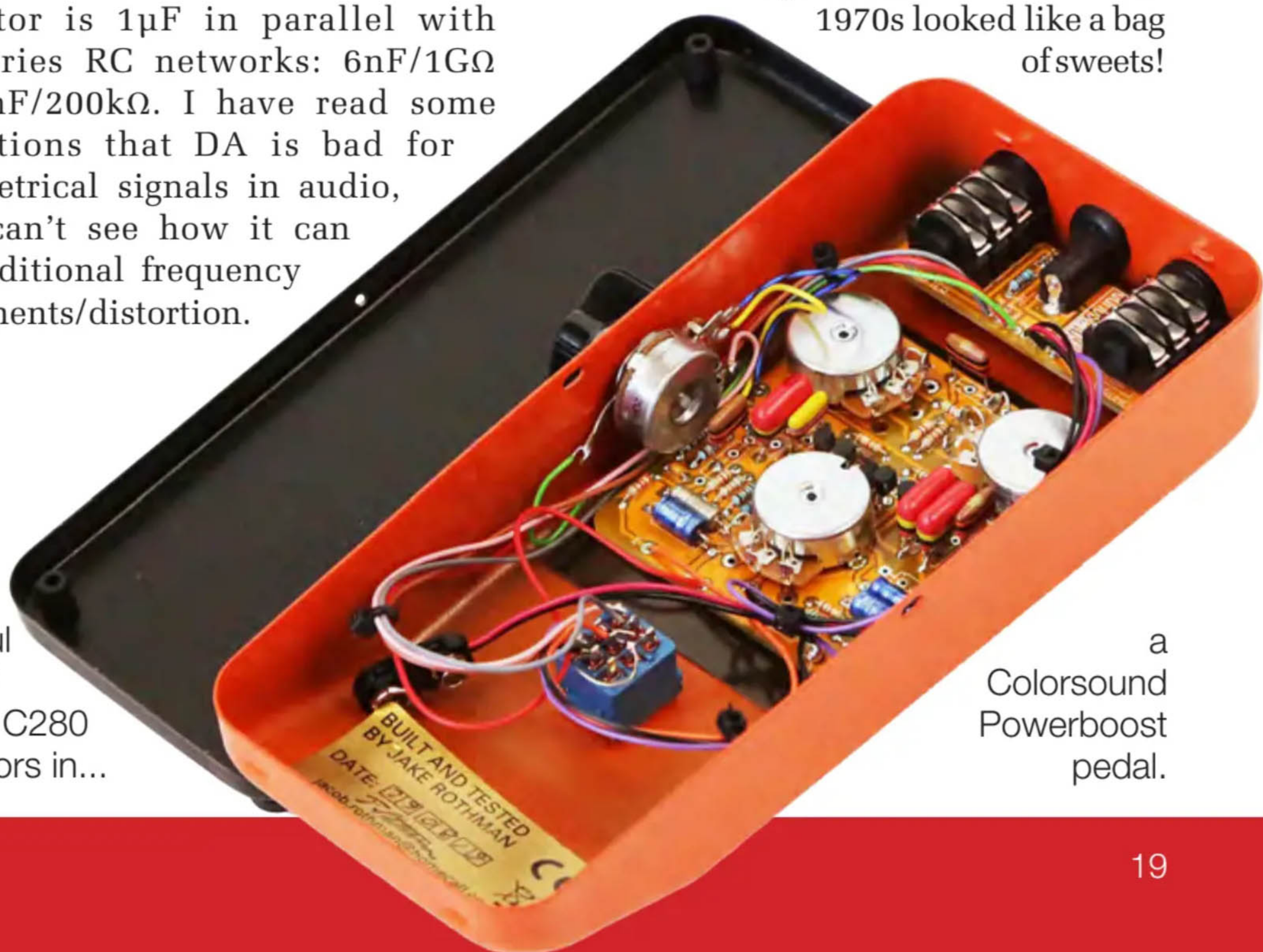
No distortion harmonics are generated from DA since it's a linear effect. On transient signals, there may be a small absorption of energy and a release in the form of a tail. It is this that makes some engineers such as Jung and Marsh believe audible distortion must occur. Maybe it could lead to bias modulation and a possible variation in THD.

It is a real concern with sample-and-hold circuits, though, as used in ADCs and analogue synthesisers. DA is bad in paper, mica, glass and Mylar/polyester capacitors. Non-polar dielectrics, such as polystyrene, polypropylene and Teflon/PTFE don't exhibit it.

A usable model for a polyester

capacitor is $1\mu\text{F}$ in parallel with two series RC networks: $6\text{nF}/1\text{G}\Omega$ and $3\text{nF}/200\text{k}\Omega$. I have read some suggestions that DA is bad for asymmetrical signals in audio, but I can't see how it can add additional frequency components/distortion.

Fig.3: colourful Phillips/Mullard C280 capacitors in...



a Colorsound Powerboost pedal.

Bass effects

The distortion from capacitors is similar to that from magnetic sources, such as balancing transformers. It's said that the distortion from magnetic tape and transformers is subjectively enhancing, and yet the similar capacitor distortion is bad, even though the capacitor distortion is an order of magnitude less.

Fig.1 shows a bad case of capacitor distortion caused by a cheap tantalum bead. This sort of problem was common in amplifiers from the 1970s, but hidden by other distortions.

In my opinion, all distortion is bad for HiFi and studio monitoring, because it leads to intermodulation and a consequent lack of clarity and muddle. However, for individual sounds within a mix, such distortion can be enhancing.

An occasional problem all passive components can suffer from is bad internal contacts. These can be identified as the 30kHz third harmonic of 10kHz by the equipment shown in Fig.2. Most manufacturers don't screen for this, so the odd rogue capacitor can be found that has bad distortion across the whole frequency range. Luckily, these are rare.

"Capacitosis"

Despite all the evidence that capacitor distortions are a subtle second-order effect, capacitor foibles have achieved a notoriety out of all proportion to their reality in the audio community, which is a strange marriage of art and technology.

I have a theory that it's psychological, that people are influenced by what they see, rather than what they hear or measure. I've christened this phenomenon "capacitosis". Maybe it's because capacitors are the most visually diverse and colourful of all electronic components. A mixed bag of surplus capacitors from the late 1970s looked like a bag of sweets!



Fig.4: the NAD 3030 amplifier – a classic, discrete and well-designed HiFi amp.

They certainly provided an initial attraction for me as a child; I remember collecting and arranging them. The Mullard C280 series “liquorice allsorts” (or “tropical fish” in the USA) capacitors were my favourite (Fig.3). They are the only capacitor where you can read the value at any angle two metres away!

The area of subjective capacitor analysis is very similar to wine tasting. I suspect that extremely objective, cost-driven engineers would go to a wine tasting with a pH meter.

I have suffered from “capacitosis” myself. This happened with a NAD 3030 amplifier (Figs.4 & 5). I put £30 worth of new pretty ‘audiophile’ capacitors into one channel and left the other as was.

Upon listening, I was convinced the modified channel was the best and then found I was accidentally listening to the other side, with the

original 5p Chinese grey electrolytics, green Mylar films and brown ceramic discs.

When I sold the unit, I hoped the recipient would notice and say something like, “I think it sounds better on the left side or something”. They never did. However, the NAD circuit did use the capacitors according to the basic audio engineering capacitor rules, ensuring the distortion was minimised. Those rules are:

1) Bigger is better

This is a rare case of bigger is almost always better, both physically and in capacitance. The linearity of a given type of capacitor depends on the voltage stress applied across the dielectric in terms of voltage per metre. Large physical size is the reason a 250V polyester capacitor produces five times less distortion than a 50V component of the same capacitance.

That is not to say you can’t get good performance with tiny capacitors – you can, but you have to be more careful to choose the right types and use them appropriately (more on choosing the type later).

A big old-fashioned capacitor will have less distortion than its micro-sized surface-mount equivalent. This is why designers always ensure (with coupling capacitors) that there is very little voltage drop at low frequencies by making the capacitance very high. Typically, one makes the -3dB point 2Hz, or 10 times less than the lowest required frequency.

One problem with this approach is that the large capacitance values take a long time for their DC level to stabilise, causing long settling times when powering up and thumps on switch-off. This is why 2Hz is the sweet spot; any higher and you risk introducing distortion; any lower, and the settling time becomes noticeable.

2) Low dielectric constant caps

The dielectric constant, K , is the amount that the dielectric multiplies the capacitance compared to a vacuum



Fig.6 (left): I love demolishing things to see what’s inside, like this electrolytic capacitor! Notice the coarse, fibrous separator paper.

Fig.7: tough side-cutters work well to open most components. The aluminium can then be unwound, away from the rubber bung.

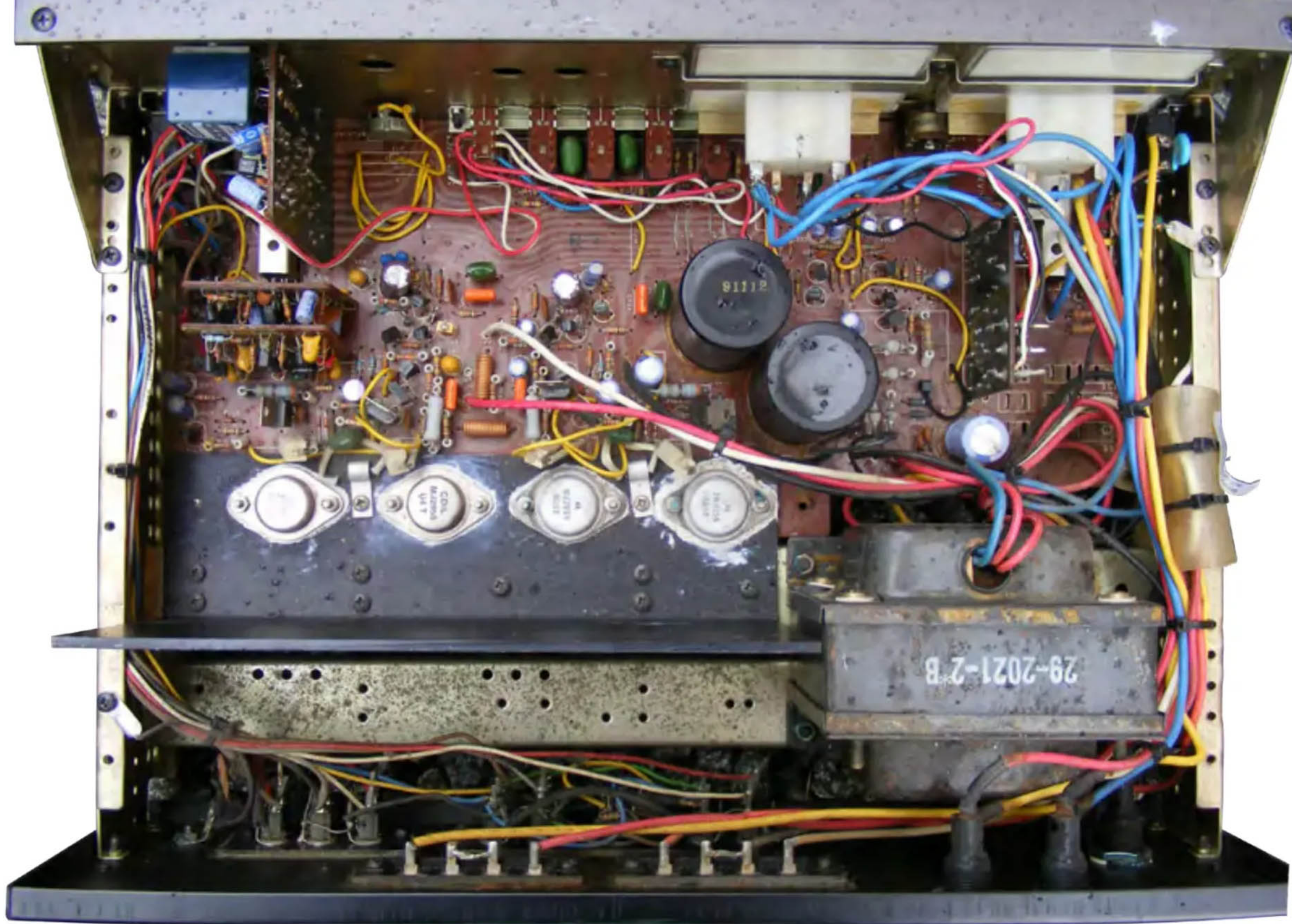


Fig.5: I swapped in expensive capacitors on one channel of this NAD 3030 amp and left the other original. I and a 27-year-old listener couldn't hear any audible difference. The design proved a high-voltage polymer electrolytic in the power amp bootstrap circuit was safe, and tantalum types in certain positions did not increase distortion.

in its place. The higher the K value, the worse the audio performance. As K increases, the smaller the capacitor is for a given capacitance. A case of bigger is better again.

Low dielectric constant ($K=2-3$) dielectrics, such as polystyrene and polypropylene, involve non-polar molecules. That is, the molecule has a uniform electric charge. Non-polar plastic films also have a negative temperature coefficient (tempco) that cancels out with the positive tempco of metal film resistors when they are combined (eg, in low-pass or high-pass filters).

Most other plastic films used in capacitors, such as polyester, have polar molecules, with a charged end.

This causes them to become aligned with the applied electric field, resulting in higher distortion.

3) Use electrolytics properly

Only use electrolytics (aluminium or tantalum) when high capacitance values are required and don't use them for audio filtering (eg, avoid using them in low-pass or high-pass filters). They are generally only used for AC coupling or supply bypassing/filtering in HiFi designs.

Any capacitor up to about $2.2\mu\text{F}$ can be replaced with a plastic film type. Wima makes a very small outline polyester series, MKS 2, which will outperform any electrolytic for distortion. The MKS2B051001N00JS in $10\mu\text{F}/50\text{V}$ is only 11mm wide and 16mm high with a 5mm lead pitch. However, it costs £3.50.

Using an electrolytic or tantalum of 470nF or even $1\mu\text{F}$ is silly, except in SMT, where tantalum capacitors will give lower distortion than X7R ceramics (SMD plastic film caps are available if a little expensive).

Small radial electrolytic capacitors are also less reliable than larger case sizes because the surface area and resulting electrolyte evaporation is proportionally higher. Fig.6 shows an unwound electrolytic, while Fig.7 shows how to open them up.

Electrolytic capacitors have a high dielectric constant, in the region of eight for aluminium oxide, 28 for tantalum oxide and 41 for niobium oxide. These are not particularly high, but the capacitance-to-size ratio is also greatly increased by surface area multiplication from foil etching and, in the case of tantalums, sintering, where tantalum powder is compressed into a porous sponge.

These K values are quite low compared

to high-K ceramics, which can be as high as 1200, but the electrolytic oxides are also very thin and have high voltage stress. This explains the relatively high distortion (non-linearity).

Bipolar or non-polarised wet aluminium electrolytics always give lower distortion than the polarised variety, even when used with a polarising voltage.

This is because the naturally oxidised cathode/negative foil (which breaks down at around 1.5V, causing extra distortion in normal polarised capacitors) is formed to the same voltage of the anode foil, so it can take the same voltage in either direction.

Some of the high-quality 'audio' electrolytics such as the Black Gate (Fig.8) are asymmetric bipolars, in that the cathode foil is formed to an intermediate (eg, 6V) rating, which gives the best characteristics of both types.

4) Be careful with tantalum caps

Solid tantalum type capacitors produce up to 10 times more distortion than wet aluminium electrolytics because of the higher K and the solid manganese dioxide, which has a non-linear resistance. So this has to be taken into account in the circuit design.

For a given CV product, eg, $100\mu\text{F} \times 10\text{V} = 1000\text{C}$, they cost six times more than wet aluminium. However, they don't dry out and, if used for low currents, such as coupling rather than decoupling, they last for decades, whereas wet electrolytics are often the first components to fail.

The leakage is 10 times lower for tantalum, and they don't need to stabilise and reform at power-on after long periods of disuse. This gives much lower clicks and thumps, and equipment can be used immediately.

Tantalum pentoxide is a more stable material than aluminium oxide and, once formed, does not degrade. Therefore, they can live without a polarising voltage and be stored for decades.

Tantalum capacitors are an important part of the sound character of the famous Neve modules. Fig.9 shows



Fig.8: special Black Gate electrolytic capacitors for audio use by Rubycon. They are so expensive that I always reuse them if their ESR is okay.

some excellent metal-cased tantalum types that Neve used.

Some engineers, like Rod Elliott (<https://sound-au.com>), disapprove of tantalums, having experienced a lot of shorts with tantalum beads. However, military and aviation engineers love them.

I suspect the antipathy comes from having to fix a lot of circuits where they have been incorrectly used. When tantalum capacitors fail, they go short circuit, unlike wet aluminium type capacitors, which slowly go open-circuit. Also, the bead type can catch fire dramatically because the manganese oxide solid electrolyte acts like an oxidising agent.

Unlike wet electrolytics, which suffer no effect from being run at full-rated voltage, the reliability of tantalums increases with a reduction in voltage. Derating, running at no more than 70% of rated voltage, is good, conservative design.

It's best to limit the surge current with solid tantalum capacitors to 1A, especially if running them at their full voltage rating. For example, if you have a tantalum bead across the output of a 15V regulator, it should 'see' a series resistance of 15Ω. This is rather a lot, and would reduce its decoupling ability. Hence my advice not to put tantalums across power rails.

Many designers did, however, because National Semiconductor used to recommend tantalum capacitors across the outputs of their regulators. Their ESR of around 0.7Ω to 2Ω is ideal for damping the

inductive output of linear regulators.

One way round the surge problem is to configure the regulator for soft starting. Another trick is to use a series decoupling resistor in the power rail. That seems rather old-fashioned today, because it reduces the regulation.

It is now possible to get surge-tested tantalum capacitors, such as the Kemet T495 series, which require no derating or current limiting.

It's difficult to get high-CV products in solid tantalum. The biggest I use is the 1000μF 10V AVX F721 for mic preamp gain control blockers (Fig.10). For SMT, they are the better choice than wet aluminium, since they are not stressed by the baking involved. Also, they seem to solder onto the board better, being shaped like resistors.

I like tantalum capacitors; used correctly, they increase reliability and reduce clicks... at a price.

5) Avoid most ceramics

Don't use high-K (X7R, Y5V, Y5U, Z5U) ceramic capacitors for coupling (or filtering). They distort much more than tantalum types. Although excellent for power rail decoupling, high-K ceramics self-modulate their capacitance with applied voltage, causing very high distortion, even if just used for coupling.

SMT electronics usually use high-K ceramics, which has led to SMT assemblies getting a bad name with regards to audio quality. However, if you can choose an NP0/C0G capacitor, it will generally perform very well. They can be quite expensive and large for a given capacitance and voltage rating combination, although they

still can be cheaper than plastic film in many cases.

For higher capacitances, especially in coupling roles, you can replace them with tantalum capacitors. For 100nF and similar values, they come in similarly sized cases.

High-K ceramics can also act as piezoelectric transducers and can pick up or emit music with very strange results. The construction of piezo tweeters is basically a lead titanate disc 'ceramic capacitor' bonded to a small paper cone.

Real differences

Generally, capacitor effects don't show up in double-blind listening trials or ABX tests. Where I have heard definite differences are in the most highly stressed audio application, the tweeter section of loudspeaker passive crossovers (Fig.11).

Just the equivalent series resistance (ESR) rating of the capacitor can make a difference. The 0.3–1.0Ω ESR of a bipolar electrolytic used in cheaper systems will make a significant difference in a 4Ω loudspeaker system, compared to the 0.01Ω ESR of a plastic film capacitor. This could mean a 25% reduction in signal level, so everybody would hear if the electrolytics were upgraded.

When I designed crossovers, I included this ESR in the circuit. When people upgraded my circuits with plastic-film capacitors, it tonally unbalanced the sound. I told them to put series resistors in to fix it.

Interestingly, special bipolar electrolytics using plain rather than etched foil are specially made for crossovers and are called low-loss, often marked LL, as shown in Fig.12. The inductor distortion in passive crossovers is much worse than the capacitors, especially if a ferrite or iron core it used (air-cored inductors are preferred for this reason, but are rather bulky).

Fig.10: tantalum capacitors don't get to very high values. This AVX F721 is rated at 1000μF and 10V, which is about the practical limit.

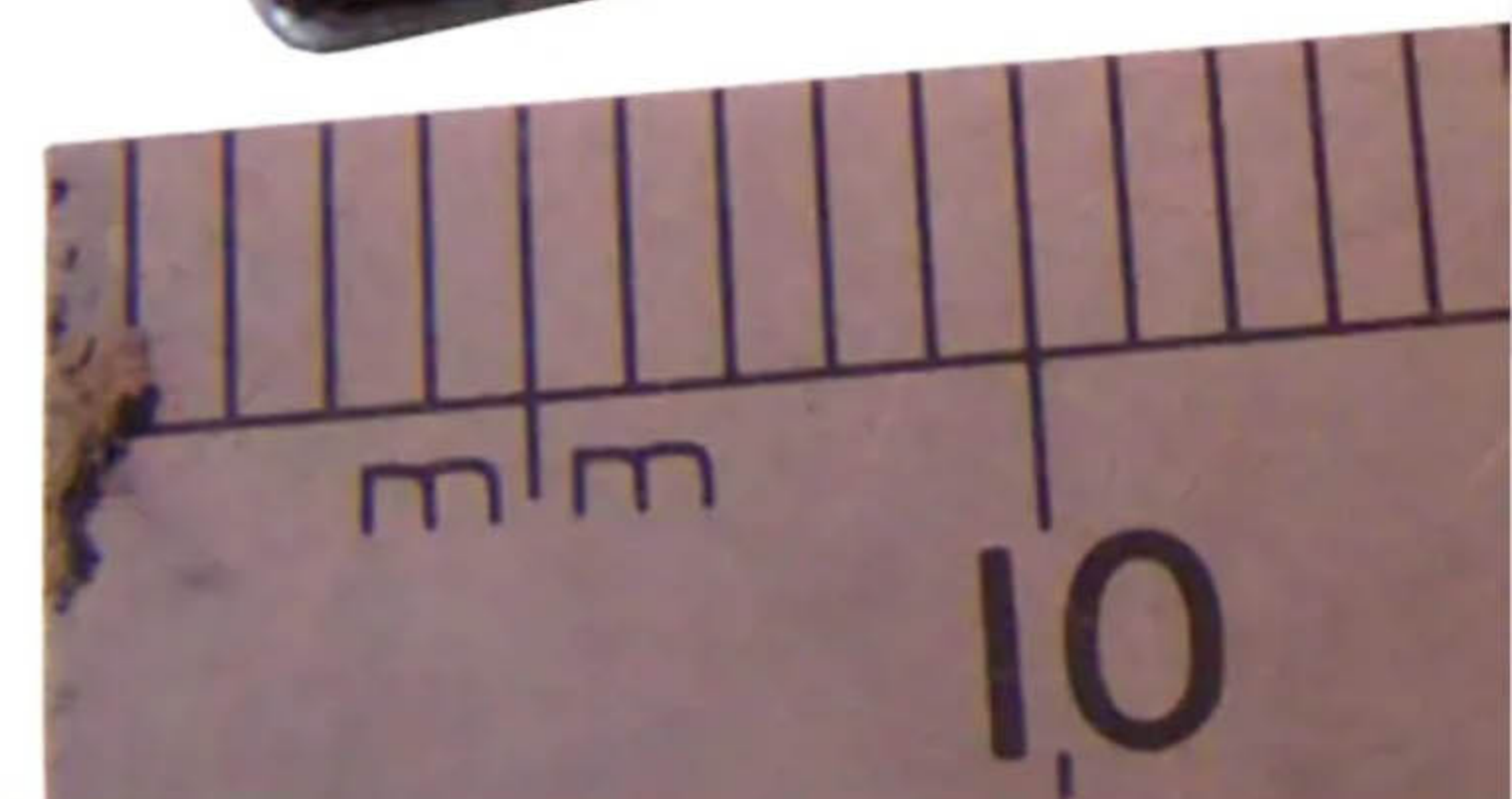
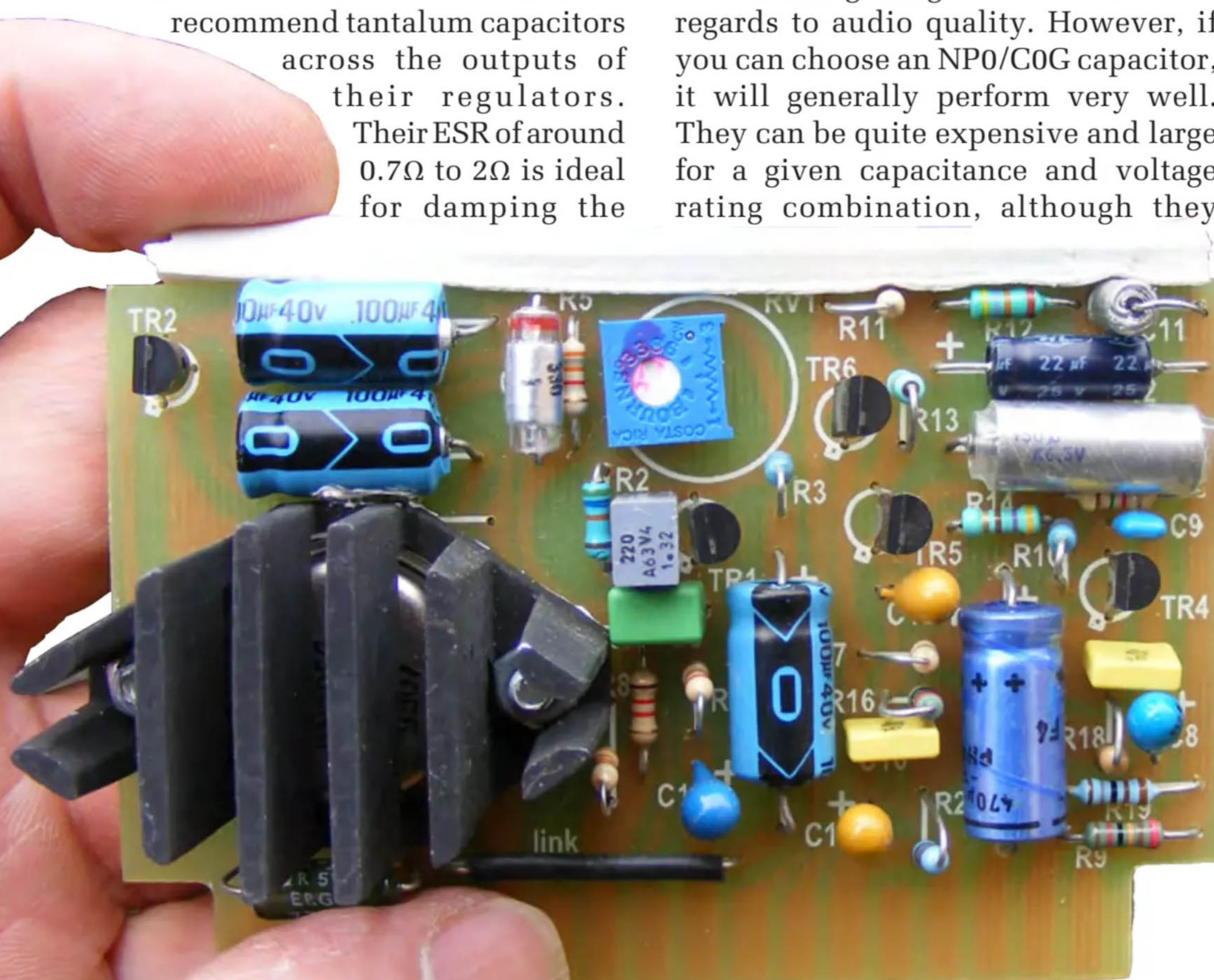


Fig.9: this Neve output module uses five tantalum bead capacitors, including one metal-cased type. I have a huge stock of RAF surplus tantalums that I use and sell to Neve and Naim module builders.



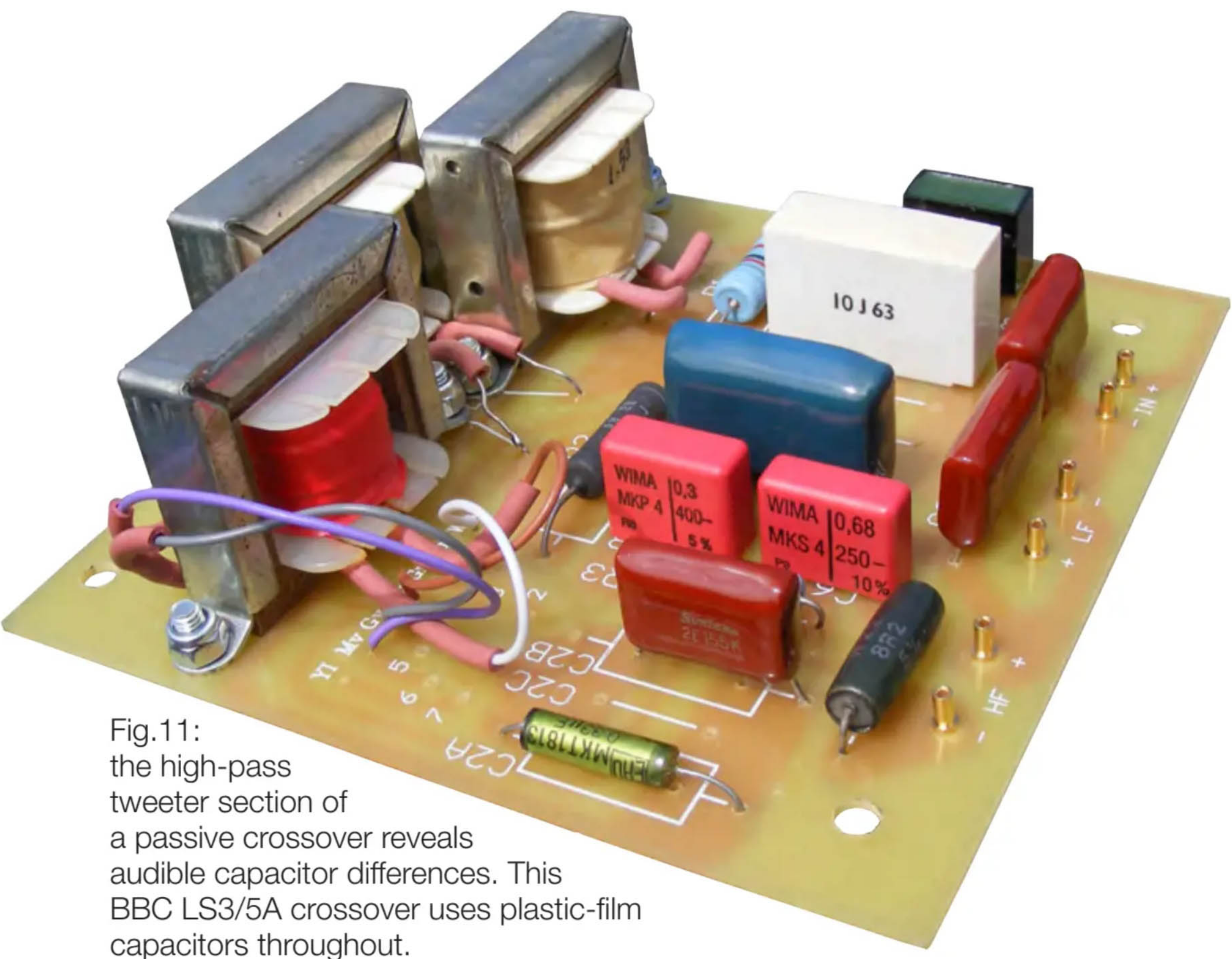


Fig.11: the high-pass tweeter section of a passive crossover reveals audible capacitor differences. This BBC LS3/5A crossover uses plastic-film capacitors throughout.

The only good capacitor is a dead no capacitor

This is a more extreme philosophy popular among some American and Japanese designers, almost ‘capacitophobic’, where all coupling capacitors are eliminated and the whole circuit is DC-coupled end-to-end. This technique gives flat distortion performance to DC, but the system may destroy itself and an expensive pair of speakers if a DC offset develops and there is no DC fault protection circuitry present.

There is no need for audio to go down to DC; doing so invites a whole load of extra problems, such as switch

clicks, scratchy pots and thumps. AC-coupling also isolates one stage from another, generally confining faults to one stage.

The need for low bias current/offset voltage audio op amps (generally single-sourced JFET input types) and DC servo circuits increases silicon costs by around five times compared to using audio bipolar devices, such as the venerable (and excellent) NE5532. This chip was designed for AC-coupled audio, so its DC performance is poor.

SILICON CHIP has designed numerous

amps and preamps with capacitor coupling that have THD+N levels down to 0.0004% and below, with very little distortion rise at either end of the frequency spectrum, and without using any exotic or overly expensive parts. So total avoidance of any capacitors in a HiFi circuit seems unwarranted.

Group delay

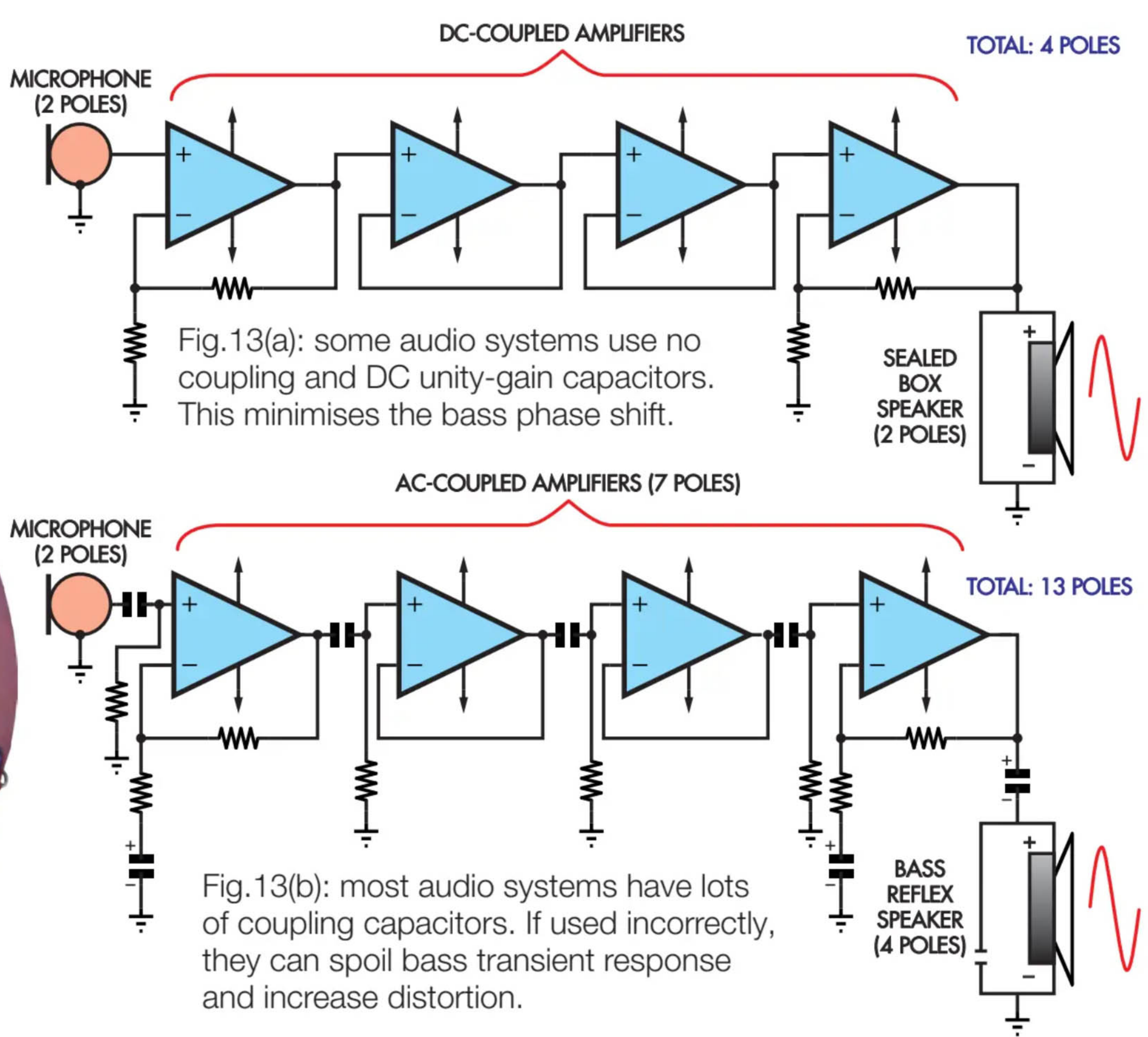
As well as acting as high-pass filters and giving rise to bass loss, coupling capacitors also have an associated phase shift. The rapid change in phase can cause a disconnect between the bass fundamental and associated transients.

How rapid the phase change is can be represented as group delay. This delay becomes noticeable on kick drums, where the ‘thump’ may seem to come late, after the ‘click’. It counter-intuitively gives a bass boost effect to male voices, making them sound ‘chesty’. This effect begins to occur at a 10 times higher frequency than the cut-off frequency.

Laurie Fincham, KEF’s principal engineer, became aware of this effect when developing the loudspeaker CUBE bass equalisers. He built two systems, one with coupling capacitors and one without, and gave a comparative demonstration at an Audio Engineering Society meeting in London.

The circuit shown in Figs.13(a) & (b) gives an example. It is impossible to remove all phase shifts and high-pass filtering from the reproduction chain. The microphone and sealed-box

Fig.12: sometimes, low-loss bipolar electrolytics like 3.3µF Callins Elcaps are used in crossovers, such as this one from a JR149.



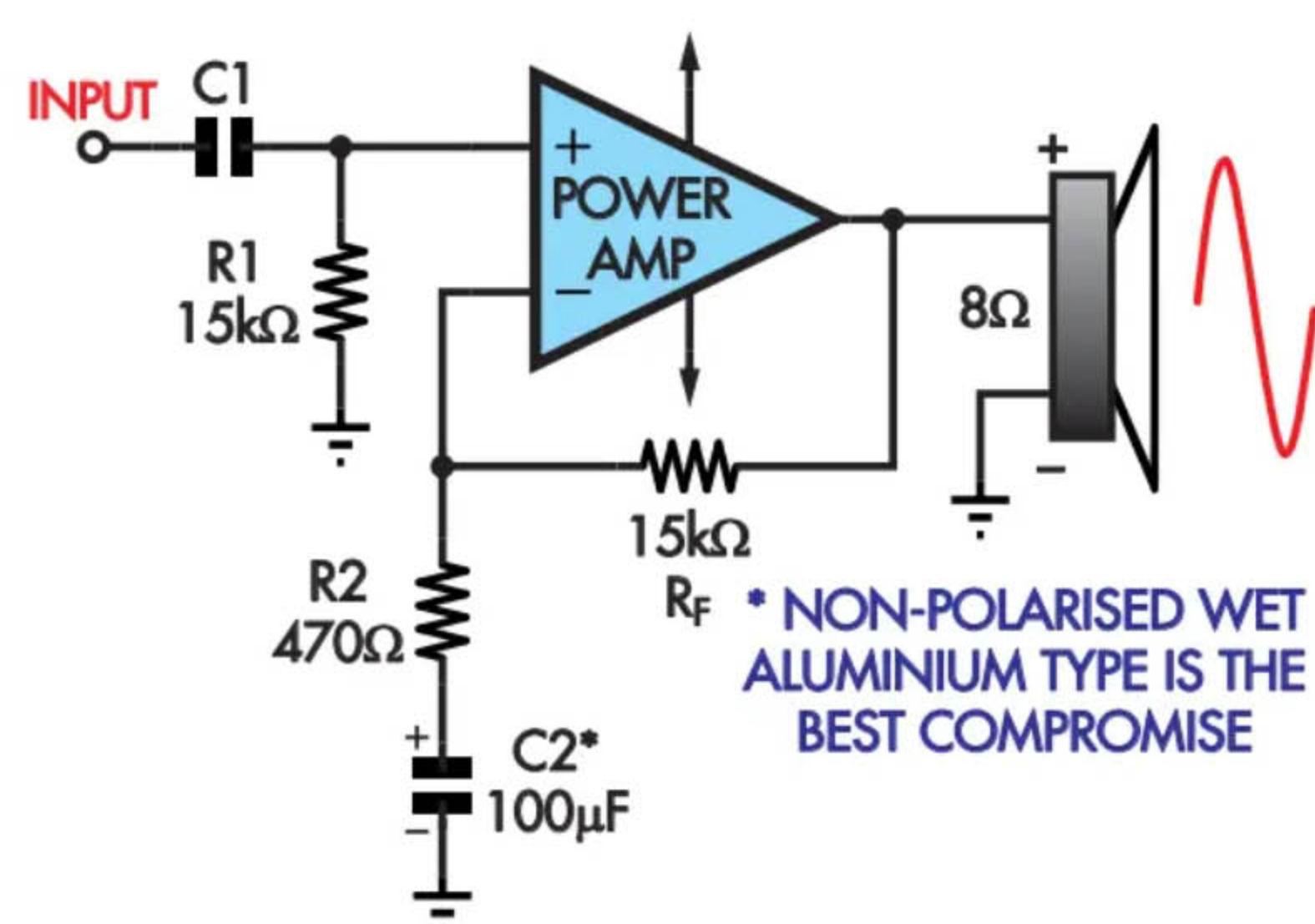


Fig.14: capacitor C2, drops the DC gain to unity. A bipolar electrolytic, like Nichicon's UVP1E101MPD (100μF 25V), gives low distortion at low cost.

speaker will have inherent high-pass filtering of second order.

However, it is possible, with a complementary bass boost to a sealed box speaker, to give a flat response down to 5Hz, where the whole system will roll off with a fourth-order slope and minimal phase shift.

Contrast this with a normal reproduction chain, with an additional seven coupling capacitors and reflex speaker, shown in Fig.13(b). Now the total roll-off is 13th order. Add in a modern sixth-order roll-off equalised reflex active loudspeaker system and the total roll-off is of the 15th order.

This leads to a large phase shift, approaching 1350°, and a consequent huge group delay, resulting in a 'warm' and 'boomy' sound. We're now accustomed to it; it is typified by Sonos speakers, Bose Wave radios, Genelec monitors and American podcasters. Most people love the effect, but it is also completely inaccurate, so the 'capacitophobic' people may be onto something.

I was surprised to read a letter raising all of this by the late mixer designer Barry Porter in HiFi News and Record Review magazine, September 1987. He also said he used non-polarised electrolytics for coupling. It seems he was ignored, like many people ahead of their time.

Edward Cherry

The lower-arm feedback capacitor used to reduce DC gain to unity in

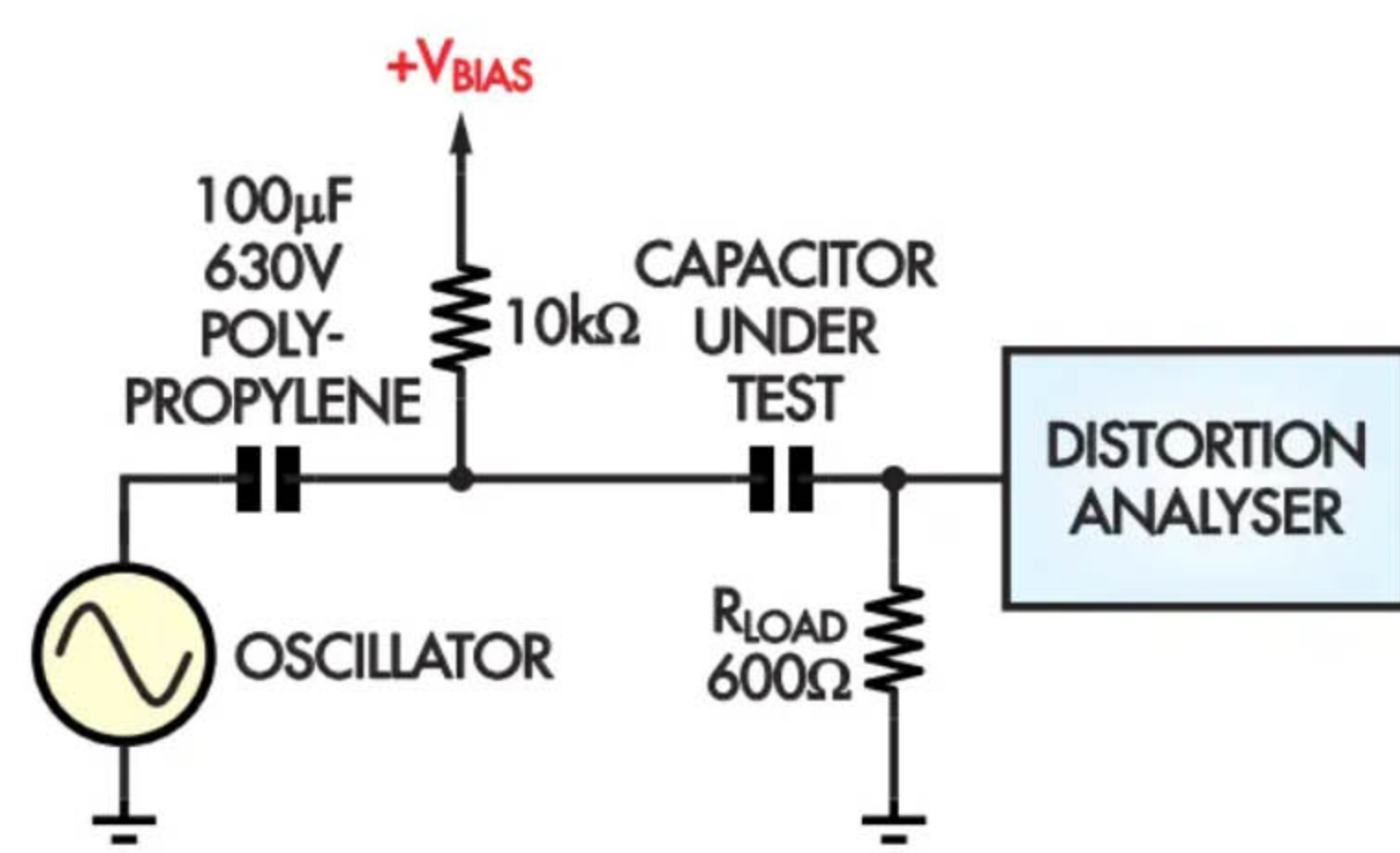


Fig.17: I used a 100μF 630V PP capacitor as a reference and for bias blocking. Its distortion was below the resolution of the AP analyser.

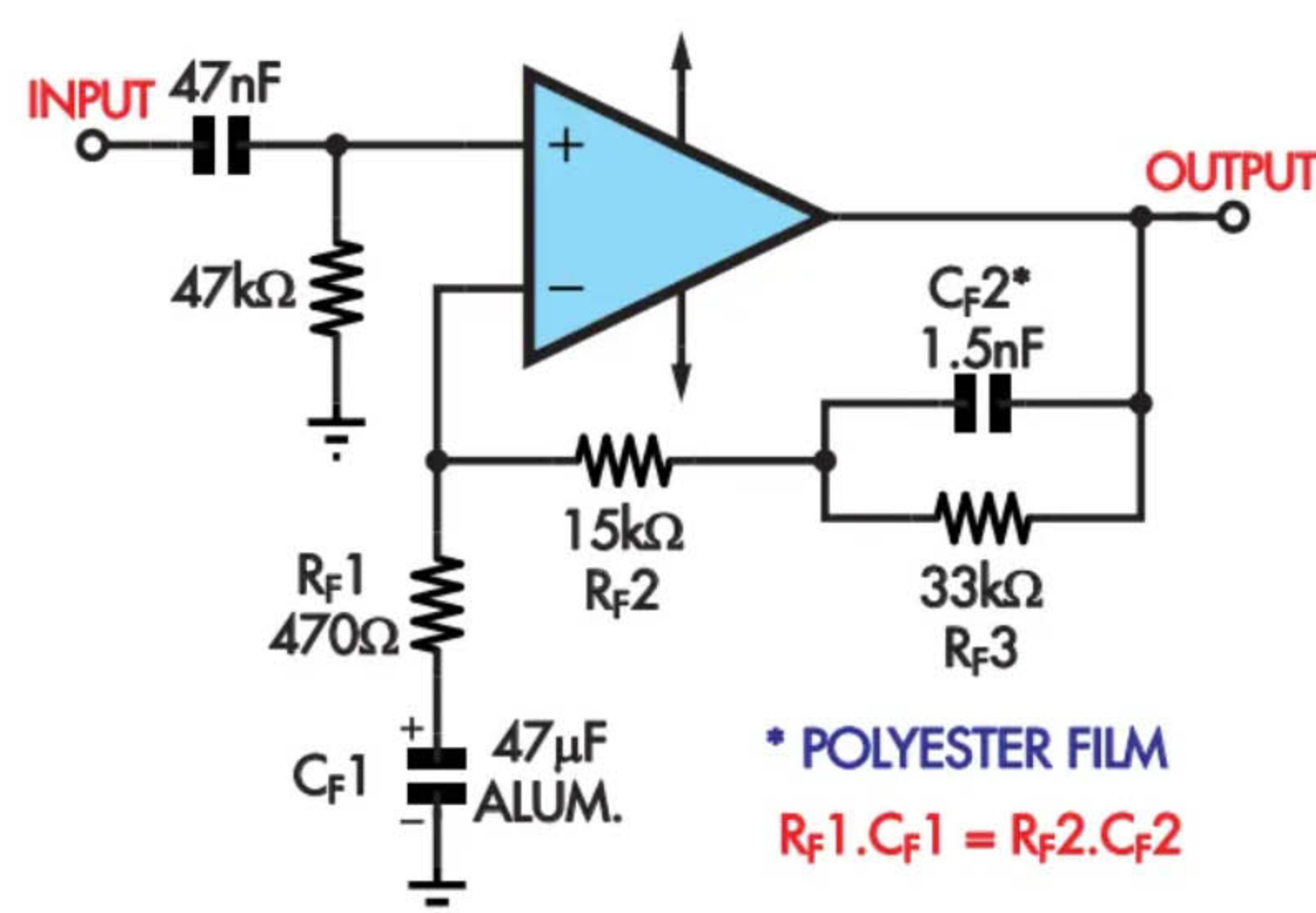


Fig.15: this shows the Cherry low-frequency compensation scheme, which minimises the signal's phase shift at low frequencies like 20Hz.

non-inverting amplifiers, such as most audio power amps shown in Fig.14 (C2 here), is critical. After all, the linearity of a feedback system can only be as good as its feedback network.

Australian power amplifier designer Edward Cherry investigated the linearity of electrolytic capacitors back in the 1970s and found them to be much better than originally believed. He also noticed that tantalum beads, which were in fashion at the time, were worse.

He used to compensate for the LF phase shift in his power amps, such as in his *ETI* May 1983 design, using the circuit shown in Fig.15. The network Cf2, Rf3 in series with the feedback resistor Rf2 reduces the negative feedback (NFB) in the region across the same range of frequencies that the attenuation of the lower feedback arm due to capacitor Cf1 boosts it.

This allows the main feedback capacitor (Cf1) to be just 47μF. Without the added compensation network, it would need to be 1000μF. He demonstrated this using a 20Hz square wave, giving the same minimal square-wave tilt for the big capacitor and his circuit with the lower value.

The voltage attenuation of the feedback network makes it possible to use lower-voltage capacitors than expected. I have

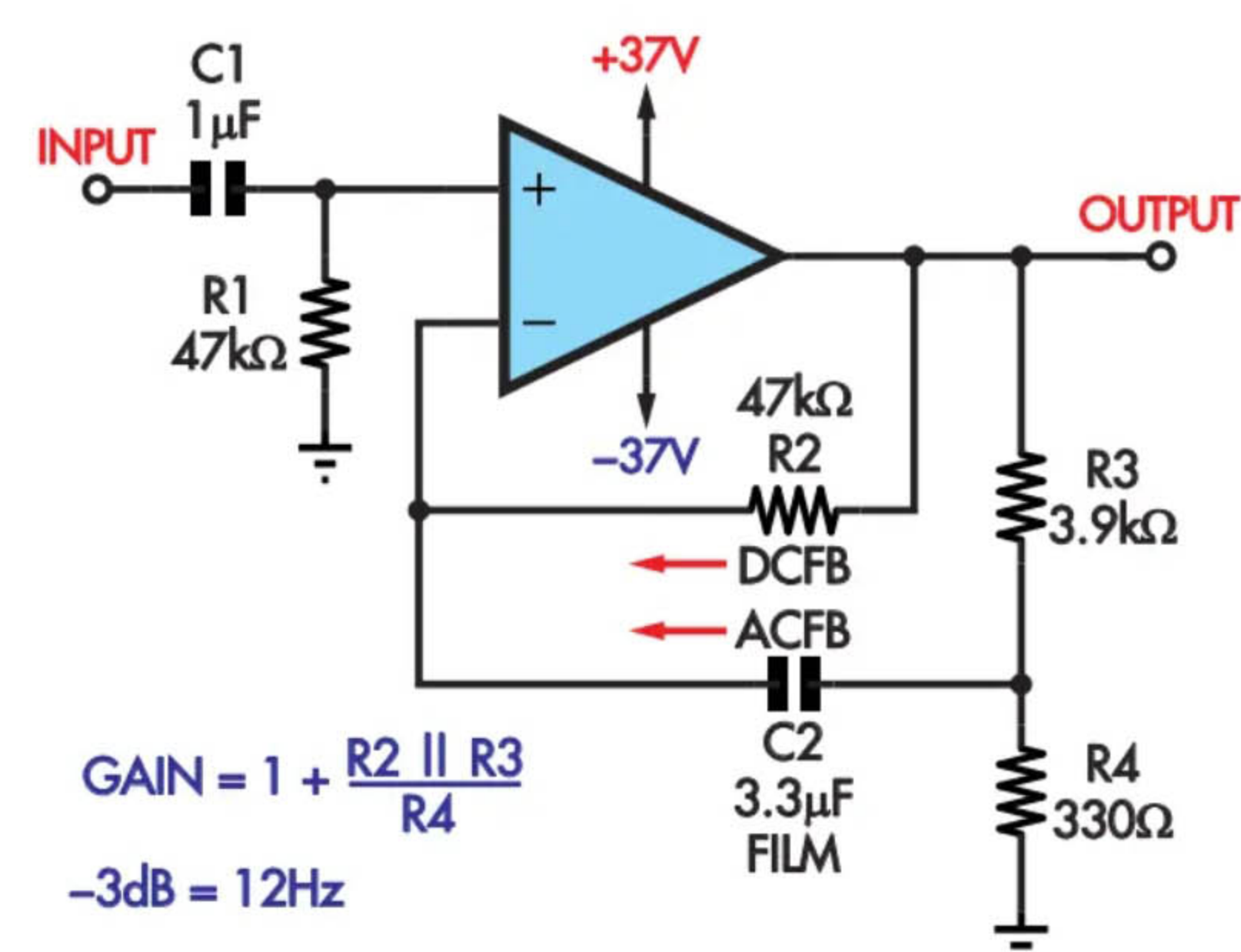


Fig.16: one way of avoiding an electrolytic capacitor in a feedback network.

seen 6.3V used with no ill effect. Due to NFB action, the maximum signal voltage across the capacitor is the same as the input voltage, typically 775mV RMS maximum.

If the amplifier latches up to one of the power rails, the feedback resistor limits the current through the capacitor to a safe value and leakage currents stop the voltage building up.

If you want to be extra safe, place 3.3V back-to-back zener diodes across C2 in Fig.14, since occasionally the charge does build up and destroy the capacitor if the amplifier is offset for some time. I have seen this with polymer electrolytics.

An alternative circuit that can be used to reduce the size of the feedback capacitor is shown in Fig.16. This allows a polyester capacitor to be used rather than electrolytic.

Both of these circuits involve the use of a high-value feedback resistor, which passes the DC bias current to the input transistor long-tailed pair. This will increase the offset voltage.

Biasing

Polarised capacitors give lower distortion with DC bias, especially solid tantalum types. Not much is required; 3–5V is usually enough.

I tested the effects of bias by connecting

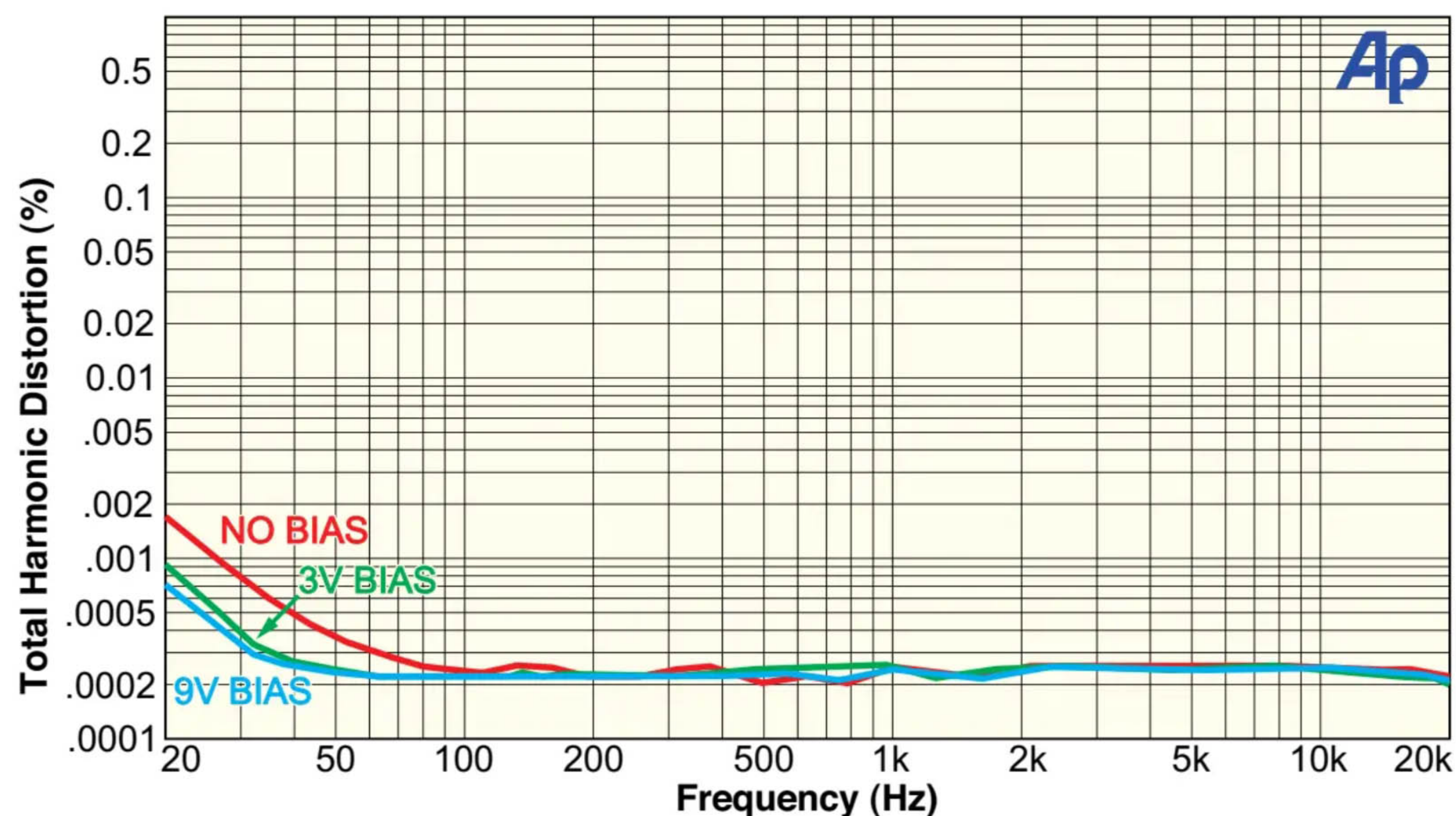


Fig.18: solid tantalum bead capacitor distortion curves, 1V RMS into 600Ω with a 100μF 10V Kemet type. No bias THD = 0.0018%, 3V = 0.0009%, 9V = 0.0007%.



Fig.19: some non-polarised electrolytics need a DC pulse around every 10 days to stop corrosion inside. Otherwise, internal gas generation can cause swelling in storage.

a very expensive 100 μ F polypropylene cap in series with the capacitor under test for DC blocking (Fig.17), which had much lower distortion than the capacitor under test. The bias was applied to the junction of the two capacitors via a 10k Ω resistor.

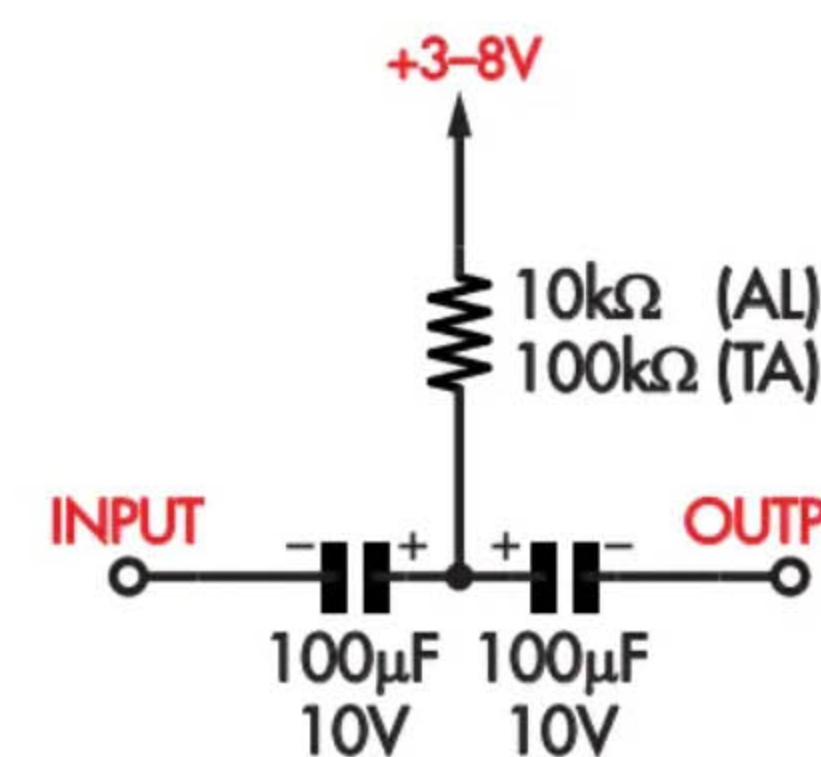
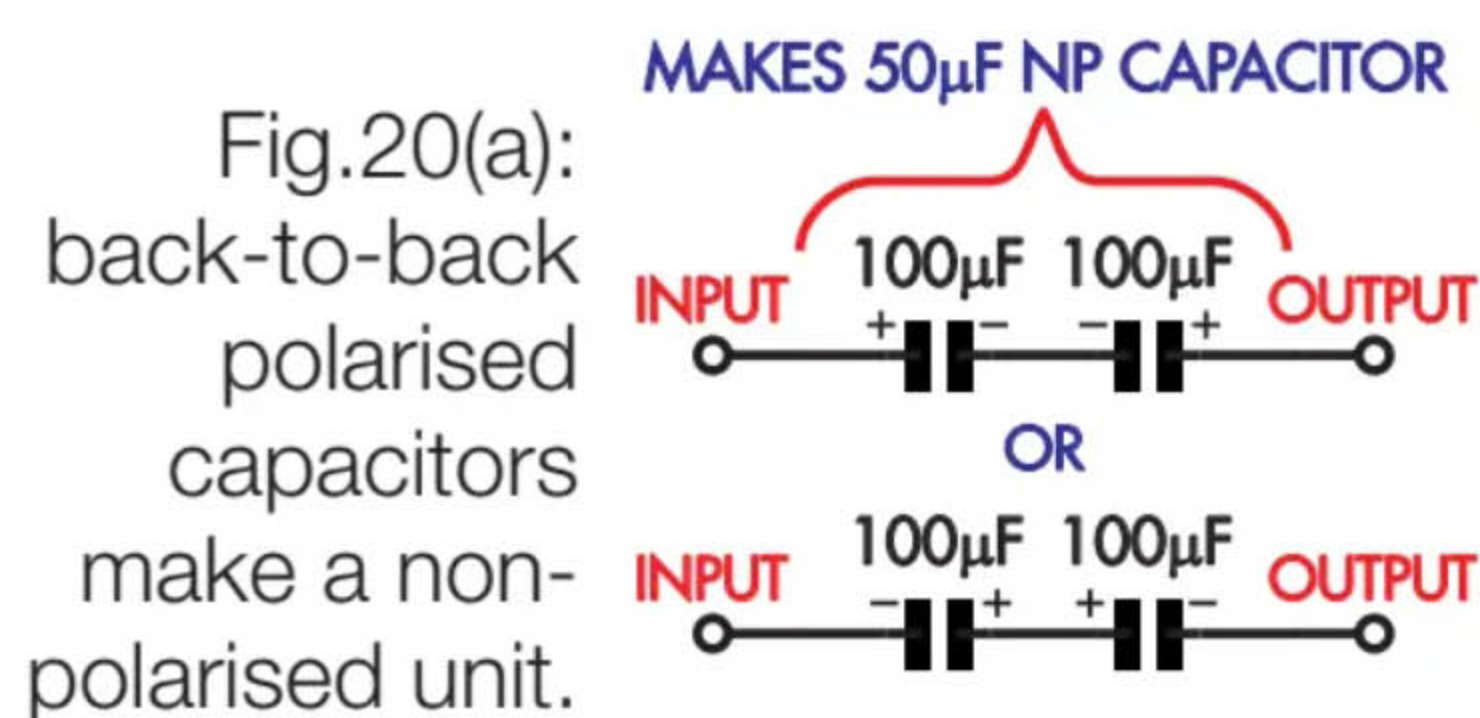
I used to worry about using polarised electrolytics in circuits with no DC bias, such as op amp systems running on dual rails. Luckily, the corrosion inhibitors in most modern electrolytes, along with the DC pulses that occur on switch-on and switch-off, are enough to prevent degradation.

Solid tantalum capacitors don't degrade without bias, but the distortion is higher (Fig.18). Wet electrolytics do suffer from prolonged storage, especially the non-polarised (NP) types. In most NP data sheets, it says to polarise one way, then the other every 250 hours. If you don't do this, the capacitors can pop in the drawer after a few years.

I had this problem with some Suntan CD71 capacitors from Rapid electronics (Fig.19).

Distortion cancellation

Putting two polarised capacitors back-to-back, as shown in Fig.20(a), can reduce distortion. It does this in several ways: first, one capacitor protects the



other against reverse polarisation; second, the second harmonic distortion (normally the biggest component of the THD) cancels out. This also occurs with the input capacitors on differential amplifier inputs.

Finally, there is a degree of self-biasing that occurs from the diode action when each capacitor is reverse-polarised. I've noticed the effect is more powerful with solid tantalum capacitors than wet electrolytics. This is because of their lower leakage and the fact that there is no additional cathode foil capacitor in series.

This self-biasing voltage takes a few cycles to build up and, with music signals, it will be unstable. It is easy to measure it with a scope or digital multimeter; it can build up to a few volts. It's much better to apply a stable bias to the centre junction, as shown in Fig.20(b).

To get best distortion cancellation, the capacitors should be matched in second-harmonic generation. Using capacitors from the same batch usually works. An example of the reduction in distortion with tantalum capacitors is shown in Fig.21.

Wet electrolytics can be very linear when used at high levels back-to-back with bias. I had to thrash these 100 μ F 35V Suntan units at the full 26V RMS output of the Audio Precision to generate the distortion also shown in Fig.21, with and without a 5V bias.

I adapted the Cherry circuit by adding a biased pair of back-to-back 100 μ F 20V tantalum bead capacitors, as shown in Fig.22. The distortion was very low

(see Fig.23). Sometimes it is desirable to obtain a long lifespan, especially at studio rack temperatures, by removing wet electrolytics.

One possible problem with this circuit is switch on/off thumps and slow ramping of the power supplies reduces it. A discharge diode for switch-off may be a good idea. This can be reverse-biased diodes across the base-emitter junction of the input transistors; a sensible, cheap precaution against transistor damage.

Next month

I have plenty more information on capacitor distortion and how to avoid it. I will continue on this topic next month, including information on the effects of physical capacitor construction on performance, a discussion of classic and special types of audio capacitors and a ranking of capacitors for audio use.

PE

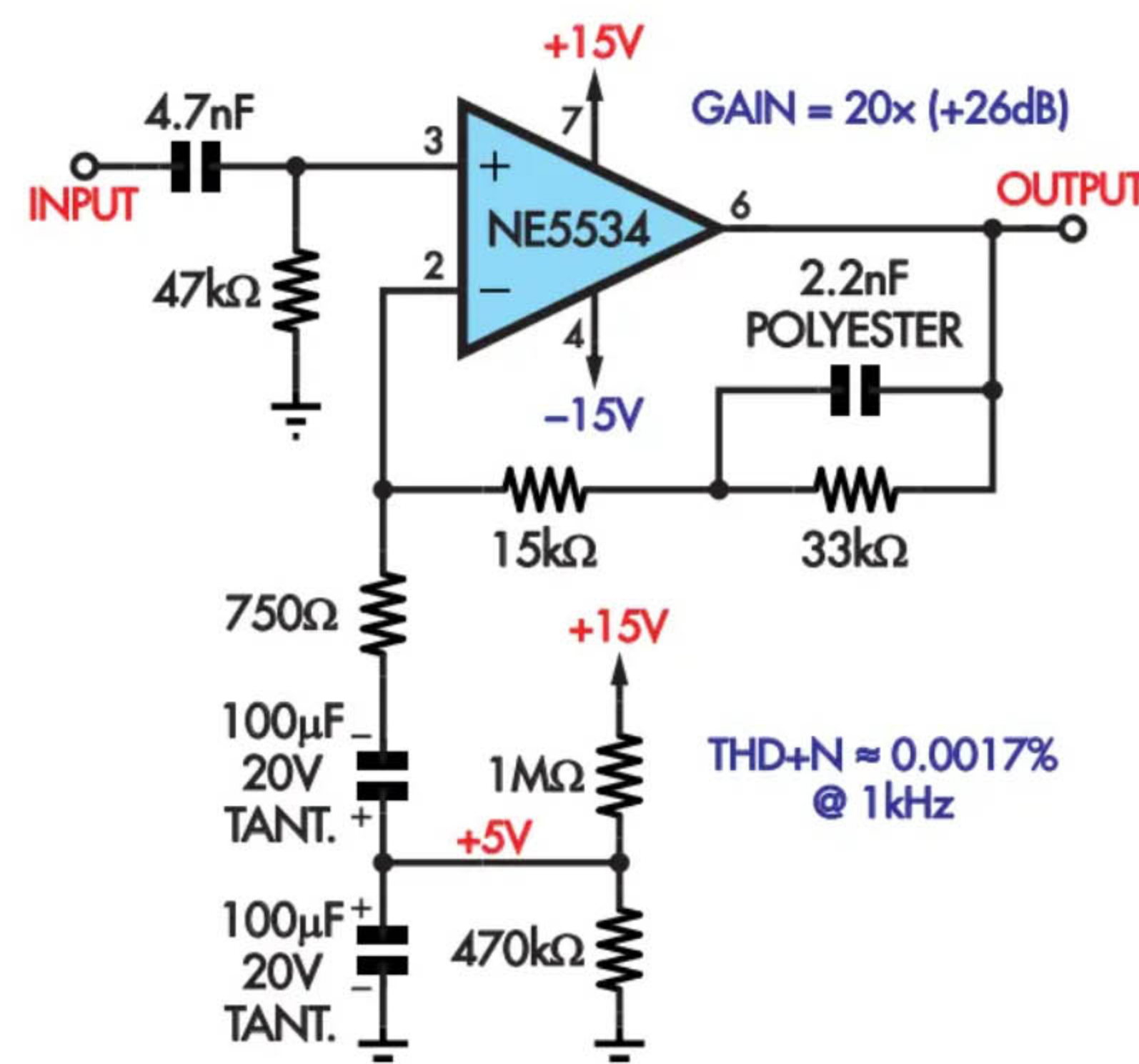


Fig.22: an adapted Cherry circuit with biased 100 μ F tantalums back-to-back.

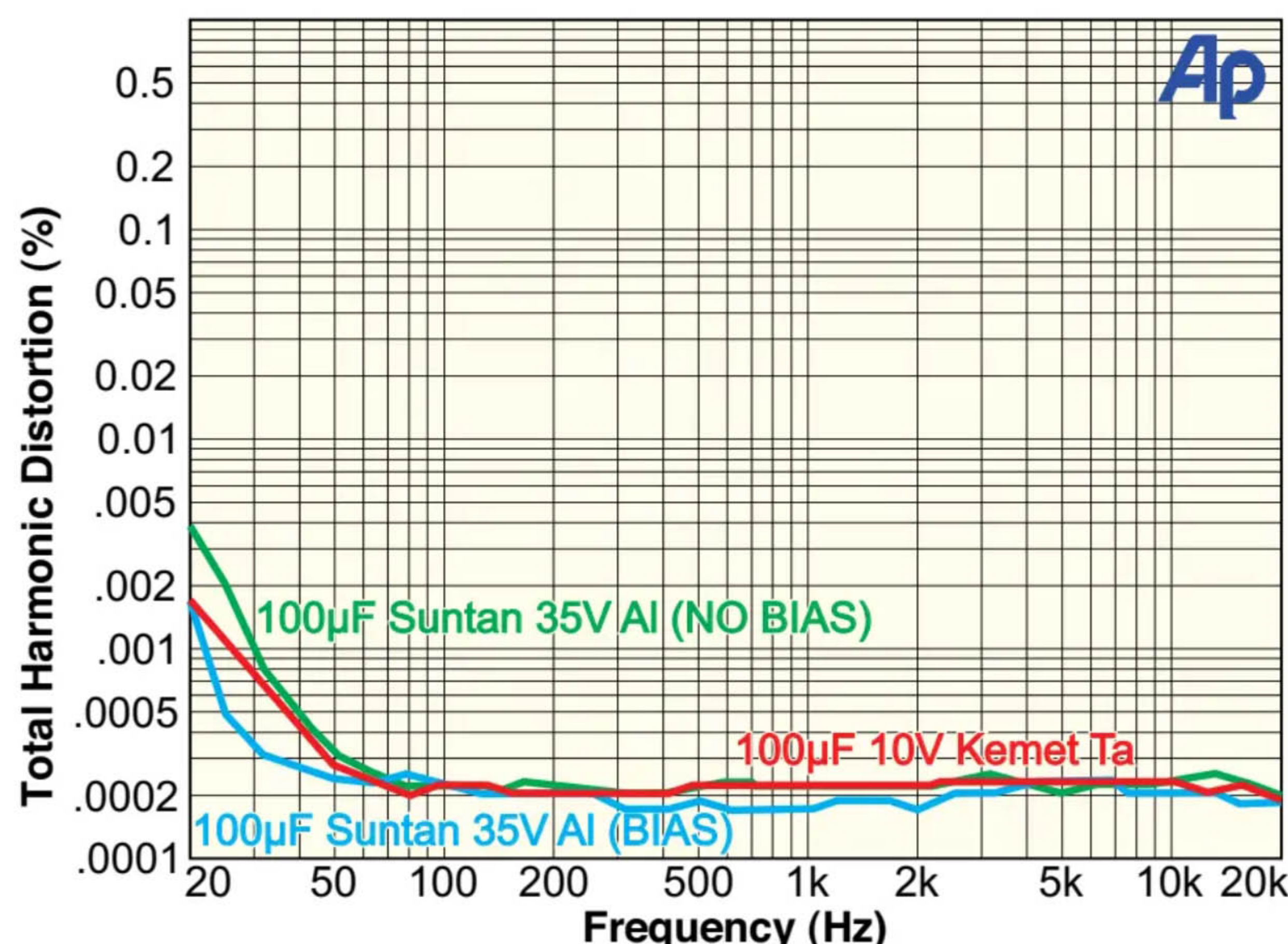


Fig.21: back-to-back tantalum beads, 4V DC bias at 4V RMS into 600 Ω , plus two 100 μ F Suntan 35V electros back-to-back.

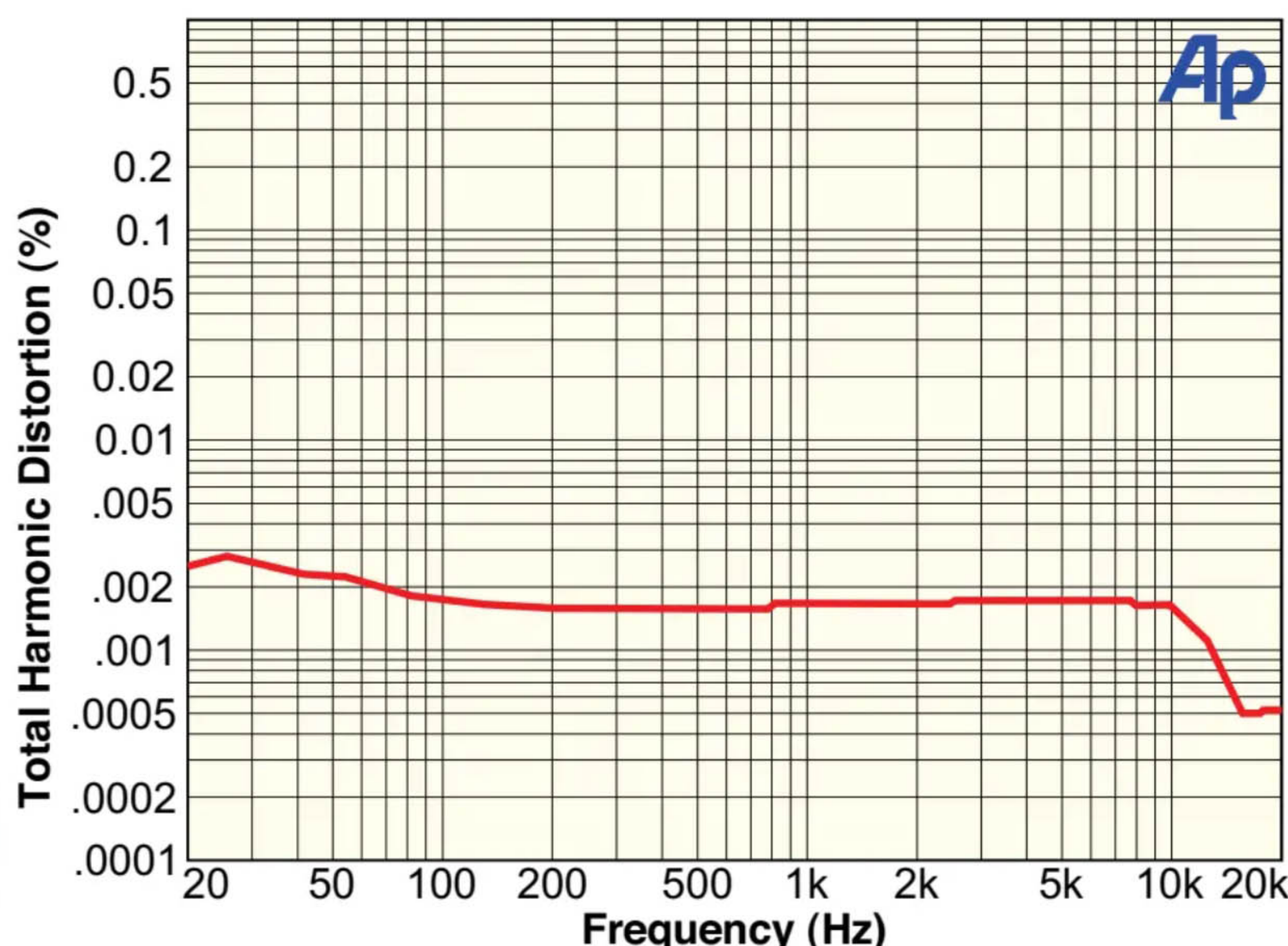


Fig.23: the distortion of Fig.22 delivering 20V RMS into 600 Ω . It has a slight LF rise but not bad for a normal aluminium cap.

Precision Electronics Part 3: difference & instrumentation amplifiers

In this third article in this series, we will further develop our precision current measuring circuit. We will consider how to sense the current if the shunt was in the positive line instead of referenced to circuit ground.

By Andrew Levido

You will recall that previously, we sensed a 0 to 1A current using a 100mΩ resistor, one side of which was connected to circuit ground. We amplified the resulting voltage by a factor of around 25 to get a ground-referenced signal of about 2.5V full-scale, which we could apply to an analog-to-digital converter (ADC) to make the measurement.

We are assuming there is a micro-controller in the circuit that can trim out much of the fixed offset and gain error, leaving us with a trimmed precision of around $\pm 0.04\%$ at 25°C with about $\pm 0.075\%$ additional error over the 0–50°C operating temperature range. We deemed this overall precision of just over 0.1% ‘good enough’ for our purposes.

In practice, we often want to sense the current in the positive leg of the circuit, as shown in Fig.1. The reason is that it is possible (sometimes even unavoidable) that the grounds of both the source and the load are connected to a common potential, such as mains Earth.

If this were to happen with the original circuit, the sense resistor would be shorted out, so measuring the current would be impossible.

Moving the shunt resistor to the positive line solves that problem but introduces another. One terminal of

the resistor is sitting at the load’s positive supply voltage (up to 20V in our example), while the other is up to 100mV higher, depending on the current through it. We are interested in amplifying only the difference in voltage between these two points, not the much larger voltage on which it is floating.

We also want the resulting amplified signal to be referenced to circuit ground so it’s within the ADC’s range, and so we don’t need to use a differential ADC to measure it.

Differential and common mode signals

In cases like this where we have two sense terminals, we refer to the voltage between them as the differential-mode voltage (V_{DM}) and the voltage at the terminals with respect to ground as the common-mode voltage (V_{CM}). This is shown diagrammatically in Fig.2.

The differential voltage of interest (V_{DM}) is ‘riding on’ the common-mode voltage (V_{CM}) that we want to ignore. For ground-referenced signals, the common-mode voltage is zero (in an ideal world, anyway).

The output of the generalised conditioning circuit block will be a voltage that is the sum of the differential-mode input (V_{DM}) multiplied by a differential-mode gain (G_{DM}), along

with the common-mode input (V_{CM}) multiplied by the common-mode gain (G_{CM}). We usually want G_{CM} to be zero (or as close to it as we can practically get) so that the unwanted common-mode voltage is rejected.

We describe the degree to which a circuit like this rejects common mode signals as the common-mode rejection ratio (CMRR). This is the ratio of the differential-mode gain to the common-mode gain ($G_{DM} \div G_{CM}$) and is usually expressed in decibels, calculated as $20\log_{10}(G_{DM} \div G_{CM})$.

You have probably guessed by now that any hope of perfectly rejecting common-mode signals (ie, achieving an infinite CMRR) is just a pipe dream. The harsh reality of electronics design means we always have to put up with something less than perfection.

Difference amplifiers

One of the most common ways to amplify a small differential signal riding on a large common-mode voltage is to use a difference amplifier like that shown in Fig.3.

Two pairs of matched resistors ($R1a = R1b$ & $R2a = R2b$) and an op amp form an amplifier with some very interesting characteristics.

This general form of difference amplifier (with separate sense and reference terminals) is a very flexible

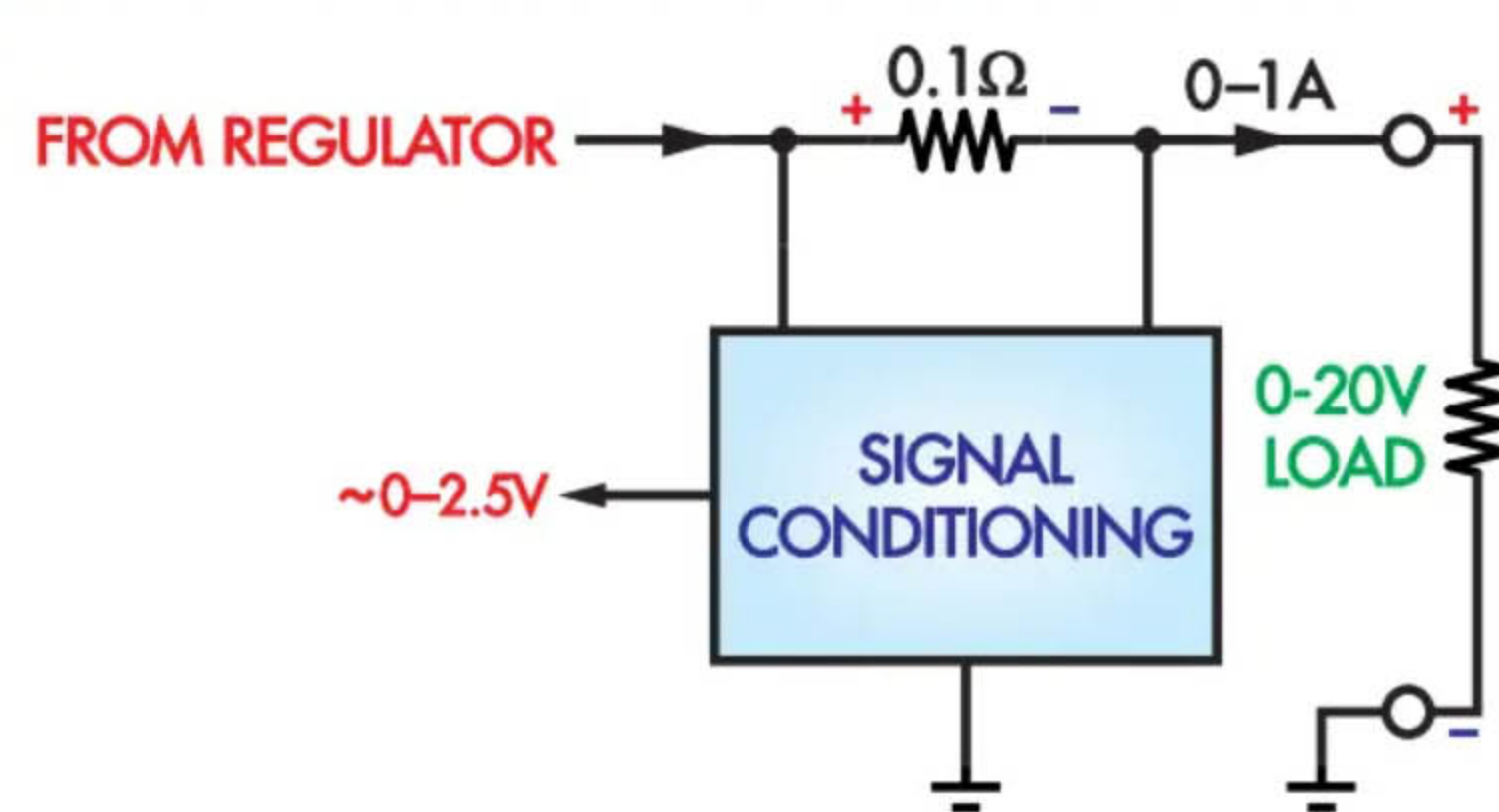


Fig.1: to measure current with a sense resistor in the positive line, we need to extract the relatively small differential signal from the larger common-mode signal.

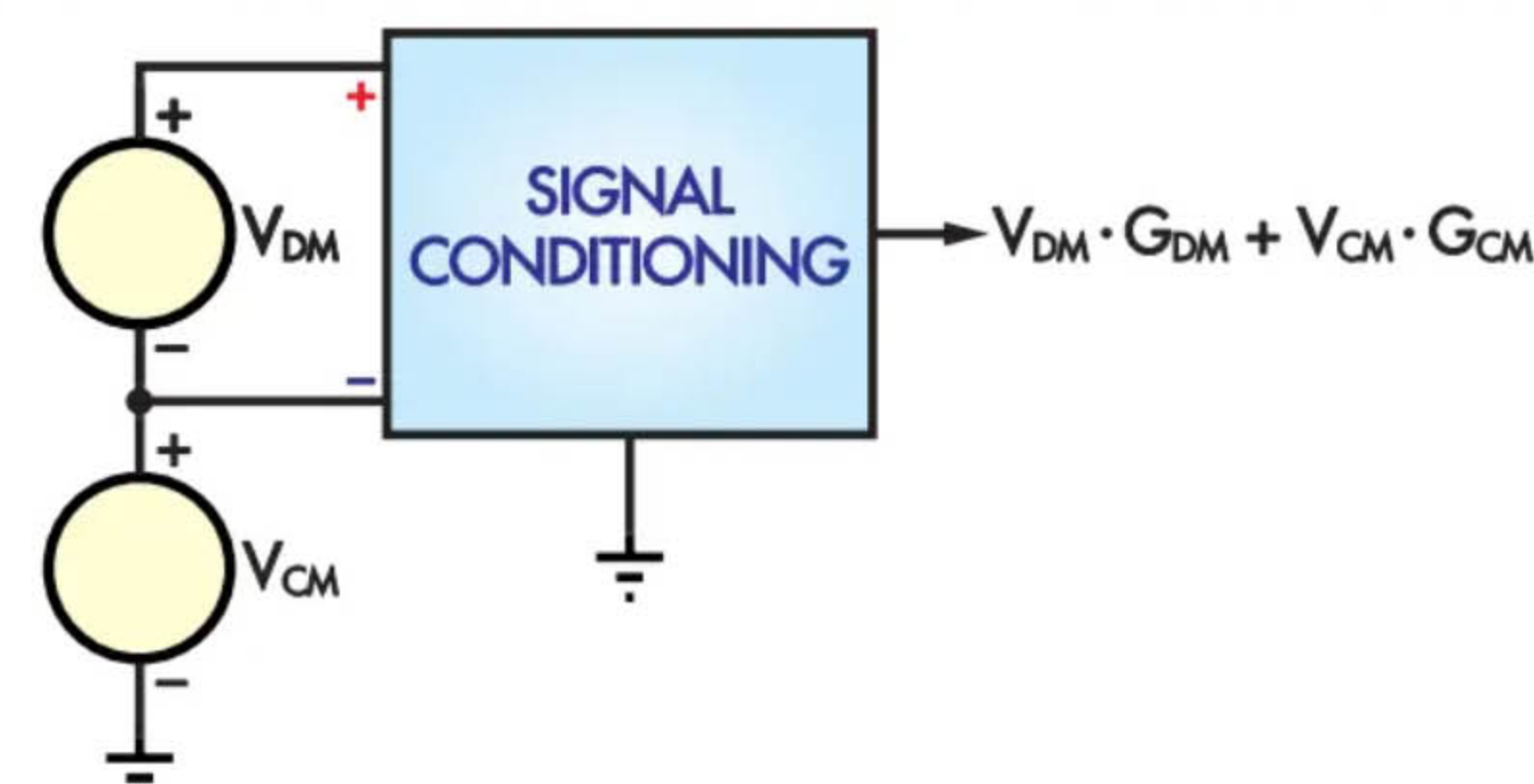


Fig.2: to achieve what we need in Fig.1, the “Signal Conditioning” box needs to amplify V_{DM} with a high gain (G_{DM}) but minimise the contribution of V_{CM} , meaning G_{CM} should be kept low.

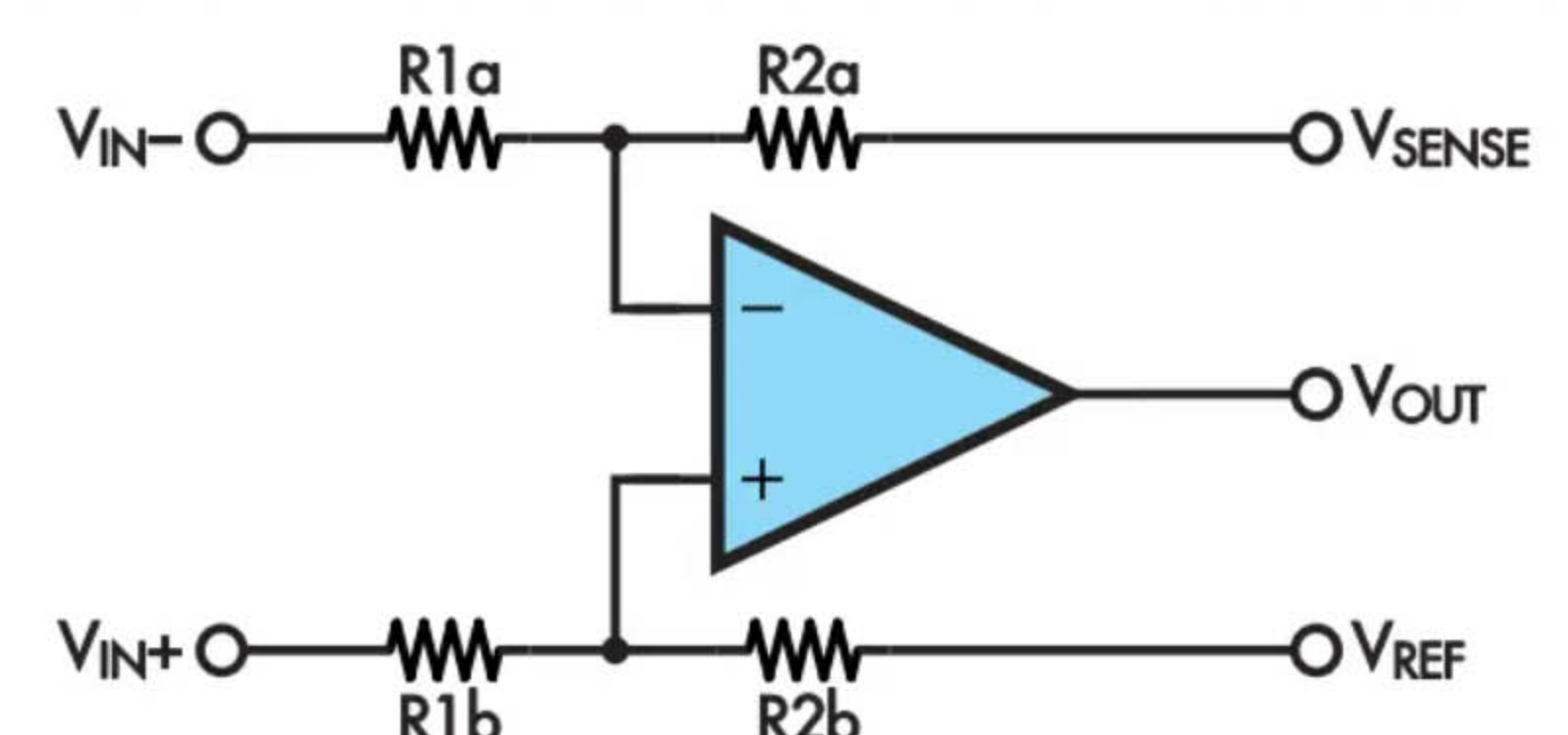


Fig.3: the classic difference amplifier using an op amp and four resistors is a very useful and flexible circuit. Usually, the value of $R1a$ is the same as $R1b$ and $R2a$ the same as $R2b$.

Difference & instrumentation amps

circuit that can be used to implement a wide variety of functions, as shown in Fig.4.

All those circuits use a difference amplifier with unity gain ($R1a = R1b = R2a = R2b$).

The terms “difference amplifier” and “differential amplifier” are often used interchangeably. I am using the former term to describe any amplifier in which the output is proportional to the difference between the input voltages.

Some sources use “differential amplifier” as the general term and “difference amplifier” to describe the specific configuration where the differential-mode gain is equal to one (eg, [pemag.au/link/ac1h](https://www.pemag.au/link/ac1h)).

To add to the confusion, the terms “differential amplifier” and “fully differential amplifier” are both used to describe op amps with complimentary positive and negative outputs. These are specialised devices are normally used to drive high-speed twisted pairs or differential input ADCs

Getting back to the device itself, connecting the sense terminal of a difference amplifier to the output and the reference terminal to ground produces the familiar configuration illustrated in Fig.5. This has a differential-mode gain of $G_{DM} = R2 \div R1$ (where $R1 = R1a = R1b$ and $R2 = R2a = R2b$).

The common-mode gain of the difference amplifier would be zero if the op amp was ideal and the resistor matching was perfect. If you build a difference amp with a typical op amp with an open loop gain of 100,000 and 1% resistors, the CMRR would be in the region of 34dB.

This means you would see about 1/50th of the common-mode voltage at the output. That would equate to 400mV in our example, almost half of the differential-mode signal we are interested in!

We can do better with matched resistors. For example, using the ACASA range of resistor arrays we used last time (matched to within 0.05%), we would have a CMRR in the order of 60dB. That still means we would see a common-mode voltage of up to 20mV at the output, which is clearly not good enough for our application.

You can buy integrated difference amplifiers with on-board laser-trimmed resistors that have CMRR values in the 80-100dB range at modest cost. If we used one of these, say with a CMRR of 90dB, the common-mode voltage at the output would be just 632µV. That is pretty good, but it still represents a 0.025% error, which will have to be added to the other errors.

There is a bigger problem, however. Off-the-shelf difference amplifiers are typically only available with gains up to about 10, with most having a gain of just one or two (we will see why a little later).

Another limitation of difference amplifiers is their relatively low input impedance, typically in the range of 10-500kΩ. That is not much of a problem with a very low impedance source such as our 100mΩ shunt, but it becomes more of a concern as the source impedance rises.

You can see from Fig.5 that any source impedance will be in series with the difference amplifier’s input resistors $R1a$ and $R1b$, potentially impacting both the gain and CMRR.

A good rule of thumb is to make sure the source impedance is lower than the input impedance of the difference amplifier by the same order of magnitude as the CMRR. So, for a difference amplifier with a 90dB CMRR and 10kΩ input resistors, the source impedance should be less than 316mΩ. Any higher than that and the CMRR will be adversely impacted.

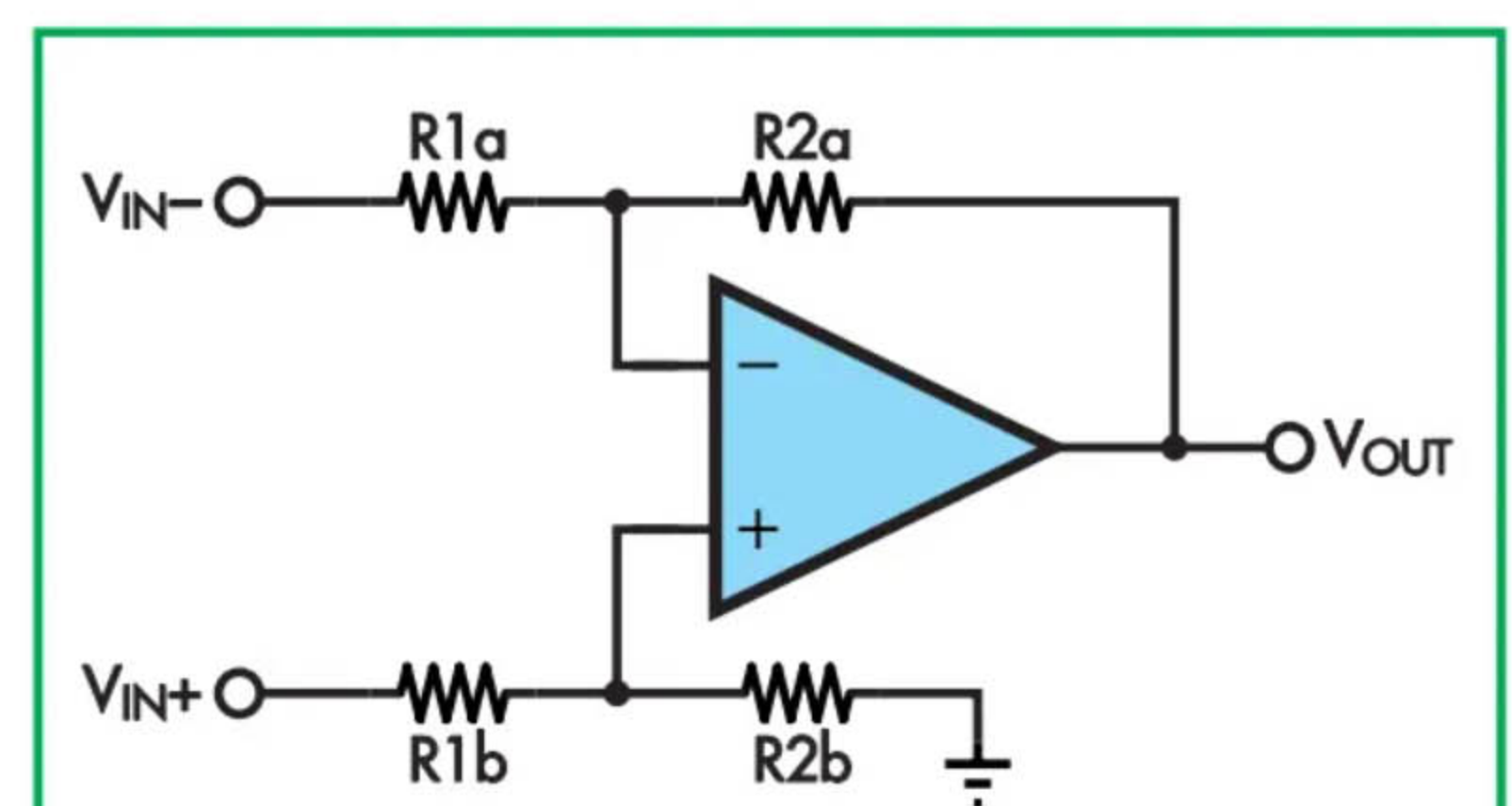


Fig.5: this configuration delivers a ground-referenced voltage proportional to the difference between the input voltages; $V_{OUT} = (R2 \div R1) \times (V_{IN+} - V_{IN-})$.

In fact, the data sheets generally specify CMRR with an input source impedance of 0Ω. That is obviously a totally unrealistic scenario – yet another reason to be wary of data sheet claims.

You might think that the CMRR would be maintained if you had equal source impedances on each input, since both input resistors would be increased by the same amount, but no such luck. The manufacturer’s laser-trimming matches the $R1/R2$ ratios in each divider, not necessarily their absolute values, which may be a bit different.

Adding the same source resistance to both inputs will likely unbalance the ratios, making the CMRR worse.

By now you might be asking why we should even bother with difference amplifiers if they have all these limitations. Apart from the flexibility we have already seen, and their role in instrumentation amplifiers that we will discuss soon, the difference amplifier excels in the area of input common-mode voltage range.

With the right resistor values, the common-mode voltage can extend well beyond the op amp’s power supply rails. Off-the-shelf devices are readily available with common mode input ranges better than $\pm 100V$. I have built discrete difference amps with common

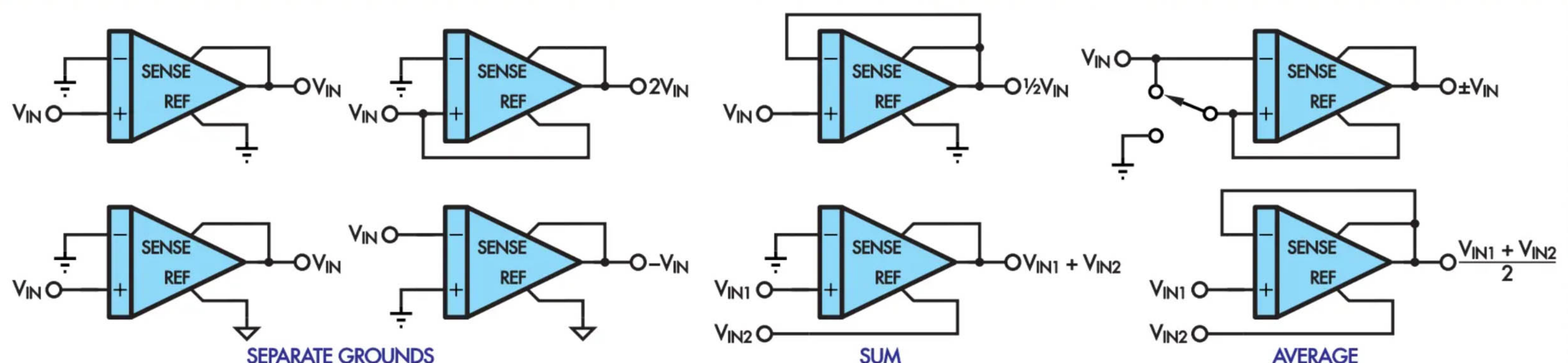


Fig.4: eight possible ways to use the Fig.3 circuit to achieve different gains, level-shift signals and even sum/average voltages.

Precision Electronics part three

Table 1: error budget for Fig.8 using an INA821

Error	Nominal Value	At Nominal 25°C		Additional error over 0-50°C (Nominal ±25°C)	
		Abs. Error	Rel. Error	Abs. Error	Rel. Error
Shunt Resistor: RESI PCSR2512 (0.5%, 15ppm/°C)	100mΩ		0.50%		0.038%
Differential Voltage due to $I \times R_{SHUNT}$	100mV	0.5mV	0.50%	0.0375mV	0.038%
InAmp: INA821 ($V_{OS} \pm 35\mu V$, 5μV/°C)	0mV	0.035mV		0.125mV	
InAmp Input Voltage total (Line 2 + Line 3)	100mV	0.535mV	0.54%	0.1625mV	0.163%
InAmp Gain Resistor R_G : RN73C2A (0.1%, 10ppm/°C)	2kΩ	2Ω	0.10%	0.5Ω	0.025%
InAmp Gain Error (0.015% ±35ppm/°C)			0.02%		0.088%
InAmp Gain (Line 5 × Line 6)	25.7	0.0296	0.12%	0.0289	0.113%
V_{OUT} DM (Line 4 × Line 7)	2.57V	0.0167V	0.65%	0.0071V	0.275%
V_{OUT} CM (20V, 120db, ±1.5db over 0-50°C)	0V	0.02mV		0.0038mV	
V_{OUT} (Line 8 + Line 9)	2.57V	0.0167V	0.65%	0.0071V	0.275%

mode voltages up to ±300V without problems (but a lot of care).

Instrumentation amplifiers

One obvious solution to the difference amplifier's input impedance problem is to add a pair of unity-gain input buffers in front of the input resistors, as shown in Fig.6. This solves the input impedance problem (at the expense of common mode voltage range), but does nothing to help us reach higher gains or achieve better CMRR.

The classic three-op-amp instrumentation amplifier (or 'inamp') shown in Fig.7 is a neat solution to the problem. The two input op amps now work to maintain the differential-mode voltage across resistor R_G . With this understanding, it is pretty easy to show that this input stage has a differential mode gain of $G_{DM} = 1 + 2 \times (R3 \div R_G)$

and a common mode gain of $G_{CM} = 1$.

We can see that with the right choice of resistor values, this input stage can improve the overall circuit's differential gain but, given it has a common-mode gain of one, it may not be as obvious how this front-end can improve the overall CMRR.

Consider a situation where we want an overall differential gain of 100 and the highest possible CMRR. Imagine the difference amplifier has a differential gain of 1 and a CMRR of 80dB. The input stage will have a differential gain of 100 and a common-mode gain of 1, giving a CMRR of 40dB. The second stage adds 80dB of additional CMRR for a total circuit CMRR of 120dB.

The instrumentation amp is effectively a gain stage with a CMRR equal to the gain, followed by a common-mode rejection stage with a differen-

tial gain of unity or thereabouts. You can now see why lots of the difference amps on the market favour CMRR over gain – they are intended for use in instrumentation amplifier applications.

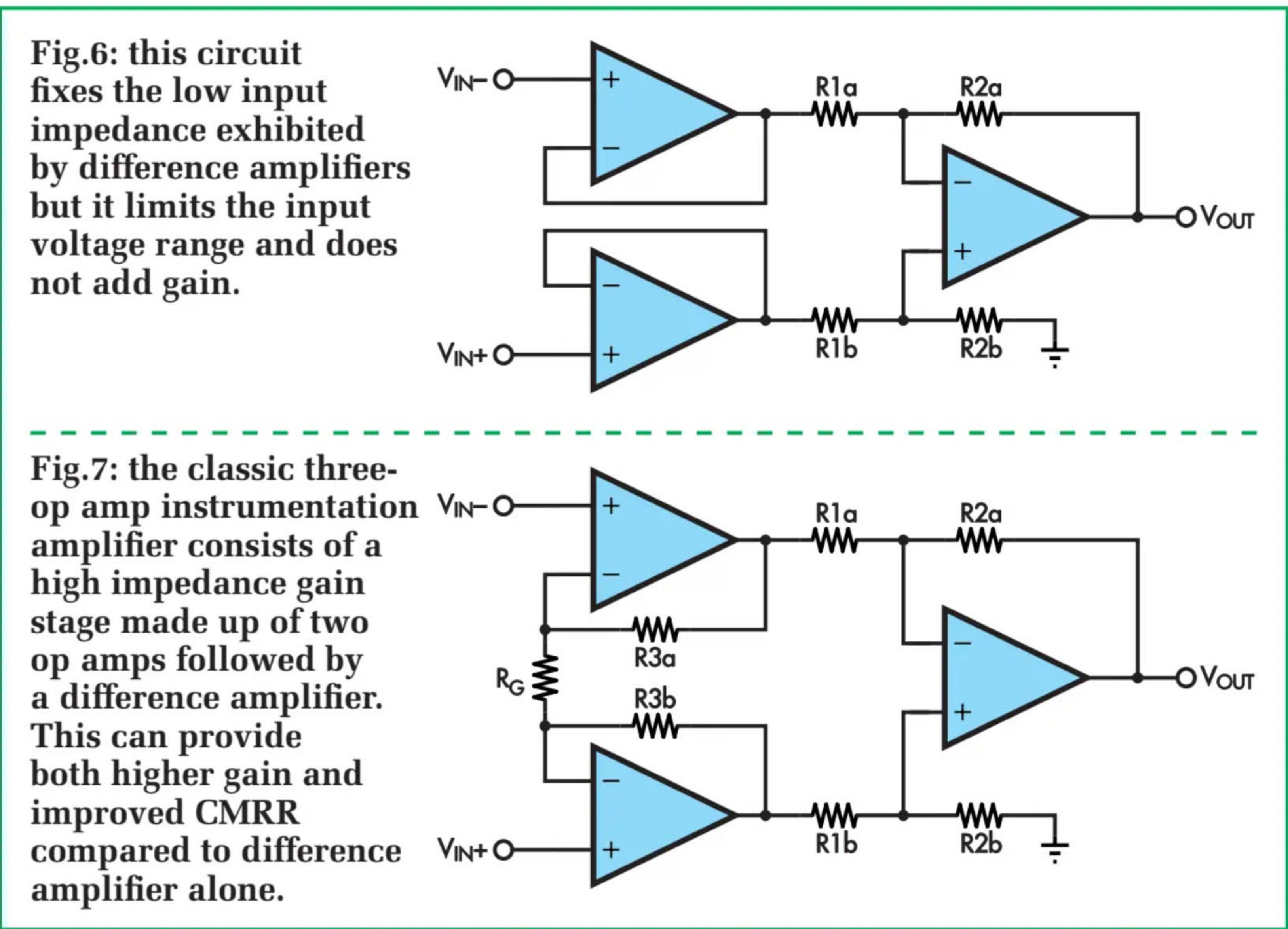
Another nice feature of the instrumentation amp is that the gain can be set by changing just one resistor, R_G . This means practical devices can have precision-trimmed matched resistors $R1a/b$, $R2a/b$ and $R3a/b$, leaving the user to provide R_G externally to set the gain. You can even switch in different resistors to change the gain or use a potentiometer to trim it.

A typical example of an off-the-shelf instrumentation amplifier is the INA821 from Texas Instruments (TI). The data sheets show it has a CMRR of 112dB for $G_{DM} = 10$ and 132dB for $G_{DM} = 100$. This suggests they are getting 92dB of CMRR from the difference amp stage (and 20dB or 40dB from the input stage).

The input impedance is 100GΩ, which should be high enough for pretty much any source impedance. The cost of the INA821 is about \$8.60 in single quantities, which is much cheaper than anything you could build yourself, given the very tight-tolerance resistor matching required.

Let's go through the process of designing the circuit of Fig.8 to compare with the ground-referenced circuit we built last time. We will build up the error budget shown in Table 1 as we go.

Fig.9 shows the internal configuration of the INA821. We need a gain of around 25, so we will choose R_G to be 2kΩ, giving a gain of 25.7. The tolerance of this resistor is not critical since we'll trim the gain, but we do care about its tempco. For this reason, I chose the RN73C2A2K0BTD from



Difference & instrumentation amps

Table 2: error budget for Fig.8 using an LT1167A instead		At Nominal 25°C		Additional error over 0-50°C (Nominal ±25°C)	
Error	Nominal Value	Abs. Error	Rel. Error	Abs. Error	Rel. Error
Shunt Resistor: RESI PCSR2512 (0.5%, 15ppm/°C)	100mΩ		0.50%		0.038%
Differential Voltage due to $I \times R_{SHUNT}$	100mV	0.5mV	0.50%	0.0375mV	0.038%
InAmp: LT1167A ($V_{OS} \pm 40\mu V$, $0.2\mu V/^\circ C$)	0mV	0.04mV		0.005mV	
InAmp Input Voltage total (Line 2 + Line 3)	100mV	0.54mV	0.54%	0.0425mV	0.043%
InAmp Gain Resistor R_G : RN73C2A (0.1%, 10ppm/°C)	2kΩ	2Ω	0.10%	0.5Ω	0.025%
InAmp Gain Error (0.02% ±10ppm/°C)			0.02%		0.025%
InAmp Gain (Line 5 × Line 6)	25.7	0.0308	0.12%	0.0129	0.050%
$V_{OUT DM}$ (Line 4 × Line 7)	2.57V	0.017V	0.66%	0.0024V	0.093%
$V_{OUT CM}$ (20V, 106db over 0-50°C)	0V	0.1002mV		0V	
V_{OUT} (Line 8 + Line 9)	2.57V	0.0171V	0.66%	0.0024V	0.093%

TE Connectivity. It has a tolerance of 0.1% and a tempco of ±10ppm/°C. The INA821's input common mode voltage range extends to within 2V of either supply rail, so we need a power supply of 22V or more on the positive side and -2V or more on the negative side.

I am going to assume we have a +24V DC supply available, since this would be the sort of input the power supply's series pass stage would need. I have already used ±5V power rails in my previous experiments, so I will power the instrumentation amplifier from +24V and -5V rails.

The INA821 has a maximum power supply voltage of 36V, so this should be fine, with a total of 29V applied (24V + 5V). It is worth noting that it is quite OK to power op amps asymmetrically like this, as long as you understand that the input common mode range and output swing will likewise be asymmetrical.

We can now complete the error budget table (Table 1). The first 8 lines of the table are similar to the previous examples, arriving at a cumulative error of 0.65% with an ad-

ditional 0.275% error over the 0°C to 50°C temperature range. Unlike the previous circuit, we now need to add the error due to the common-mode signal making its way through to the output. With a gain of 25.7, we can estimate the CMRR to be 120dB based on 92dB for the difference amp stage plus $20\log_{10}(25.7) = 28\text{dB}$ for the input stage.

With a common-mode voltage of 20V, we will therefore see 20μV at the output. That's insignificant compared to the 16mV of error due to the differential-mode stage.

The change in CMRR with temperature is a bit harder to estimate. TI provides graphs that show the temperature variation of CMRR for five sample devices at gains of one and ten. From these, I have taken a value ±1.5dB over 0°C to 50°C. It is a bit of a guesstimation, but it does not matter since the overall level of common-mode feedthrough is so low as to make this figure irrelevant.

The net result is shown therefore shown at the bottom of Table 1. The worst-case untrimmed error at 25°C is ±0.65%, just a little worse than the ±0.55% error for the ground-referenced circuit. In both cases, most of this error is the 0.5% shunt resistor tolerance.

Unfortunately, the circuit's perfor-

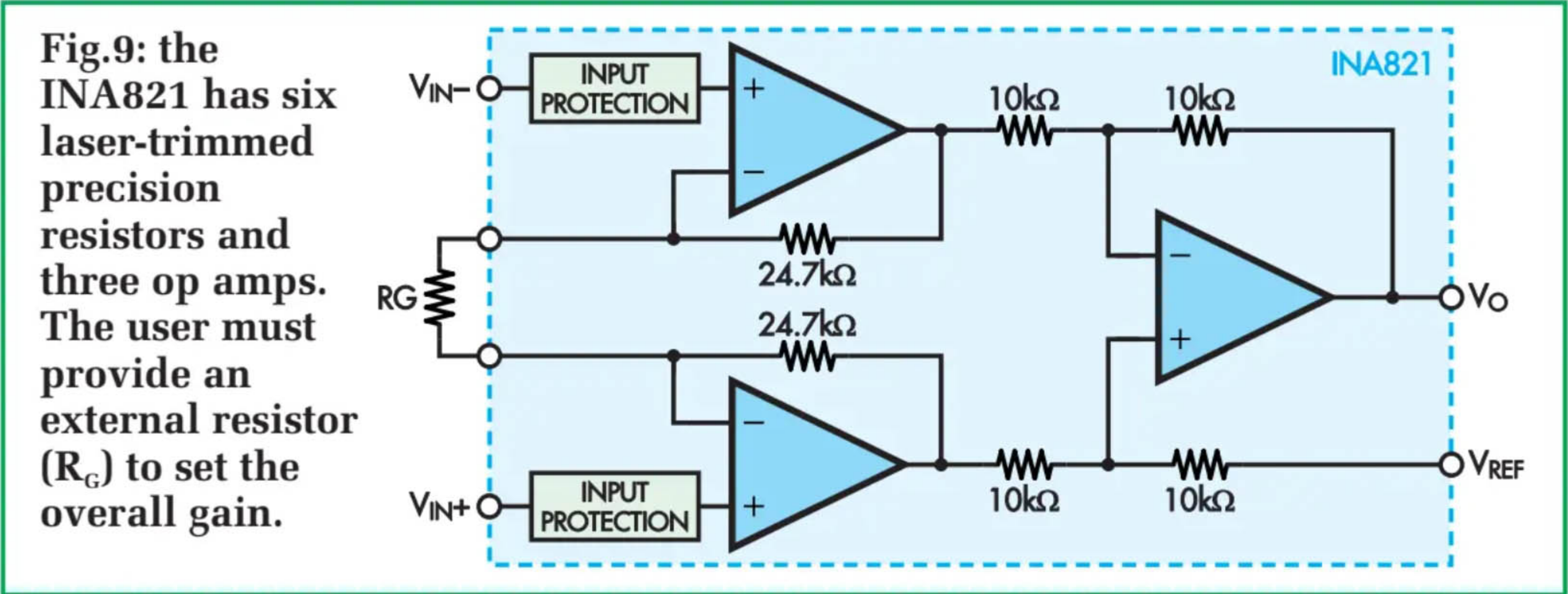
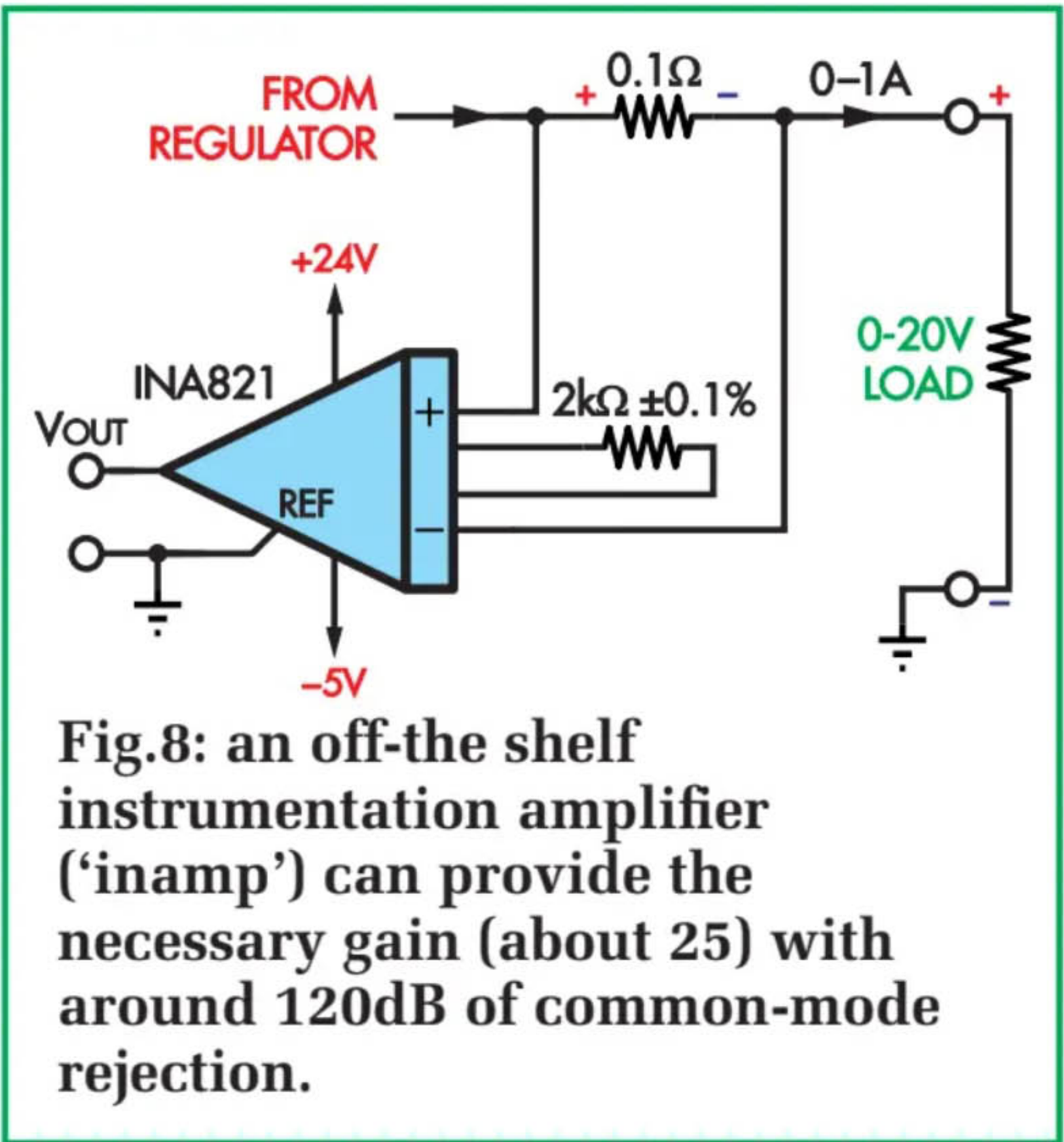
mance over the temperature range is not great. We are seeing ±0.275% error, with two major contributors: the instrumentation amplifier's input offset voltage drift and its gain drift. The LTC2057-based circuit was much better at 0.075%, as we would expect from an auto-zero op amp.

Doing better – but at a price

I wanted to see if we could improve on this, so I looked for a 'better' instrumentation amp. The LT1167A fits the bill. Its input offset voltage at 25°C is similar to the INA821, but its offset drift is 25 times better at 0.2μV/°C. Its gain drift with temperature is also better at ±10ppm/°C, compared with the ±35ppm/°C. Table 2 shows the error budget for this version of the circuit.

As an aside, it's a good idea to create these error budget tables in a program like Excel or LibreOffice Calc. I set up the formulas so that I can easily try new parts and have the whole table recalculate automatically.

Compared to the INA821, the new circuit shows a similar error at nominal temperature of ±0.66%, but an error over the temperature range three times better at 0.093%. So, we should use this device, right? Well, the LT1167A costs \$30 each in one-off quantities,



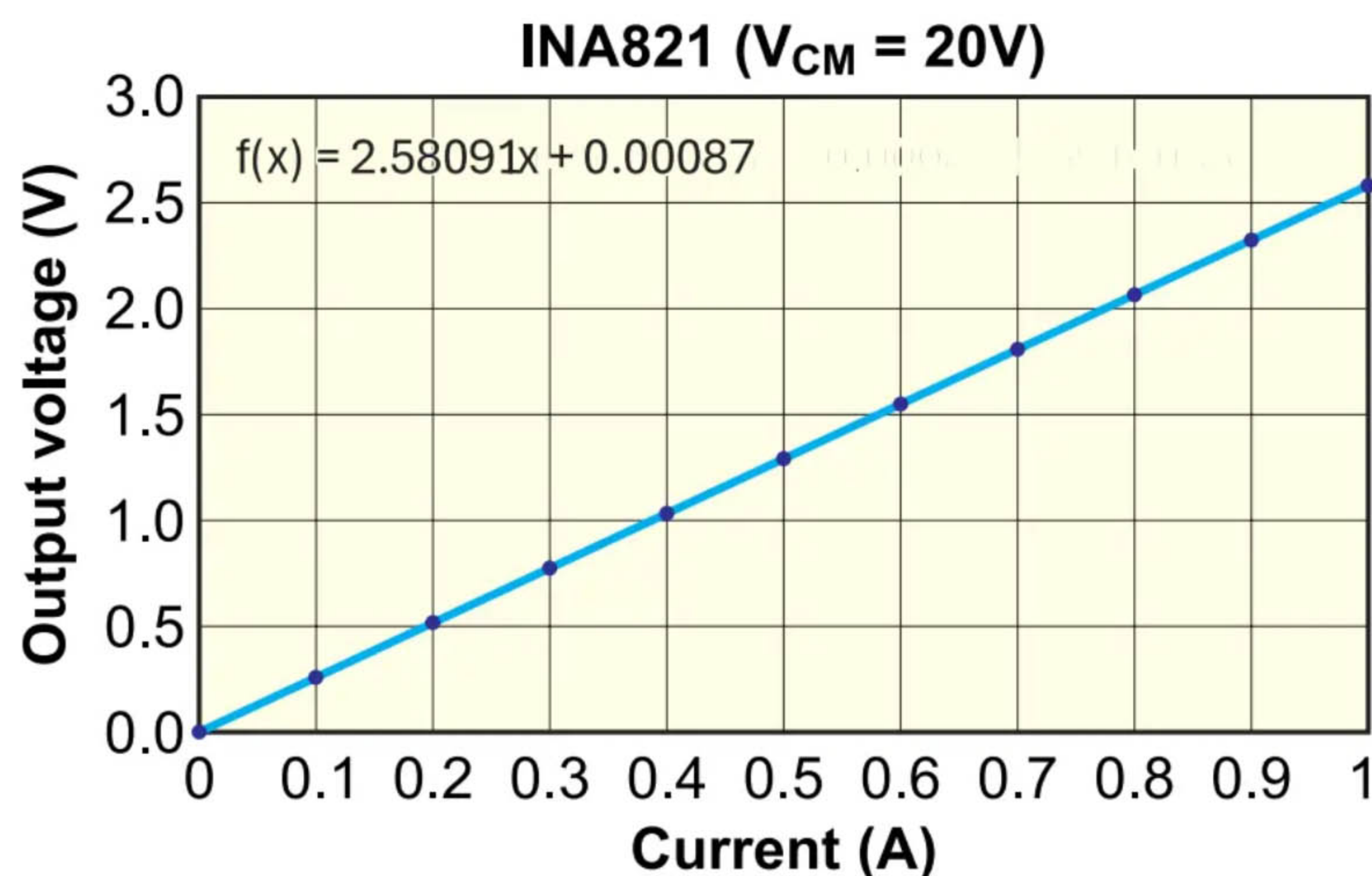


Fig.10: the measured untrimmed data for the INA821-based circuit shows about 0.3% gain error; most of this is due to the shunt resistor tolerance.

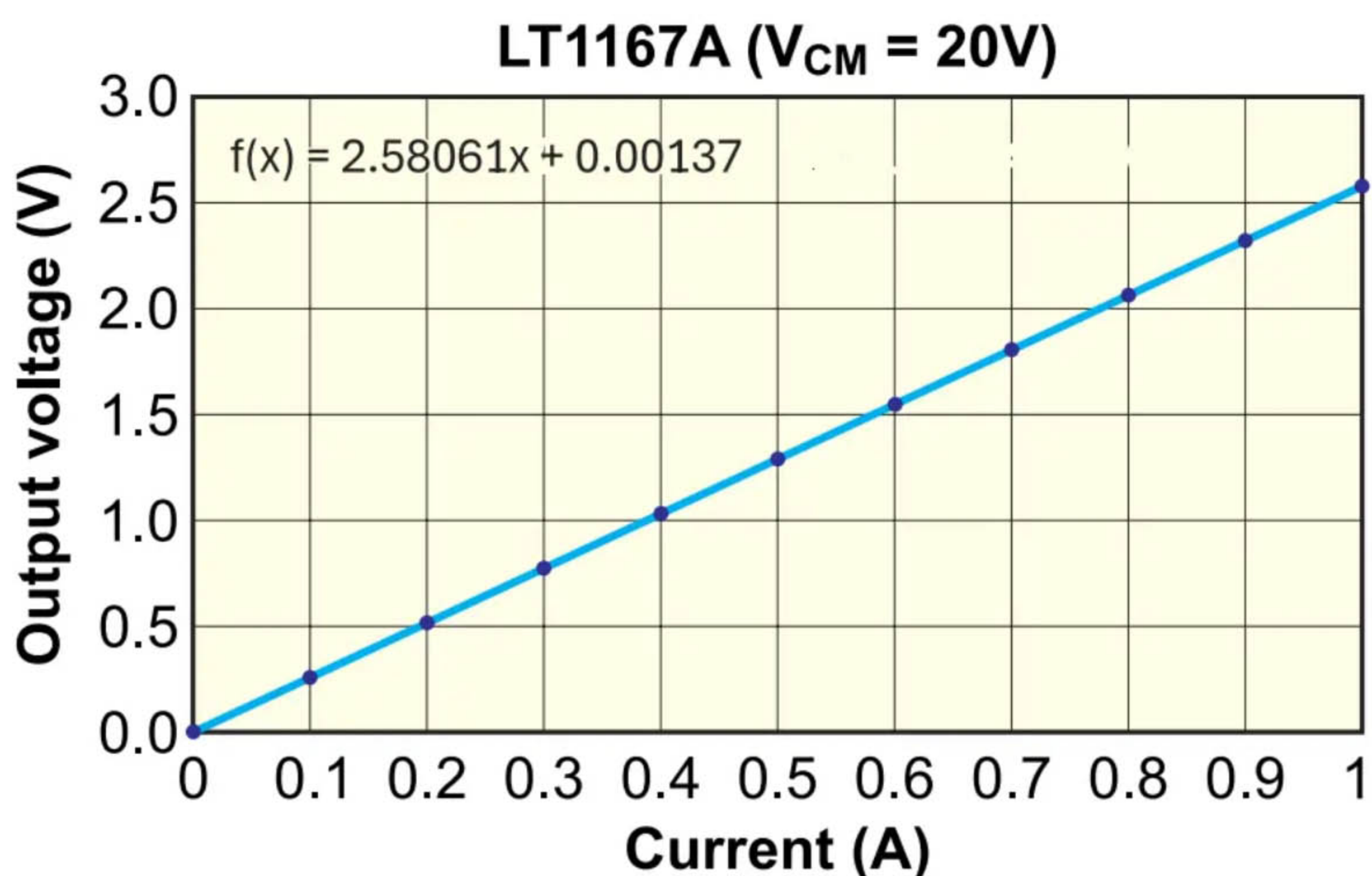


Fig.11: the untrimmed data for the LT1167A-based circuit shows the same 0.3% gain error as Fig.10 but has more offset error. It should perform better over the temperature range.

so we would want to be certain there was no alternative.

It should however come as no surprise that precision components that are at the very extremes of performance will be costly. The manufacturers know full well that if there are no or few alternatives, you will have to pay up.

Test results

I spared no expense and tested both devices. I built the circuits and measured the input current vs output voltage characteristics with both zero and the full common-mode voltage of 20V.

The results for $V_{CM} = 20V$ are shown in Tables 3 and 4, and plotted in Figs.10 & 11.

For the INA821, the untrimmed errors range from 0.01% at zero current to around 0.33% at 1A. The results were a little better with zero common mode voltage. As expected, this is better than our error budget's 0.65% worst-case estimate. The errors increase steadily with the magnitude

of current, suggesting a gain error is the main contributor.

The graphed results and line of best fit shows this to be the case. The offset correction we need to apply is very low (around 250 μV) and the gain error is about 0.3% (the measured gain is about 0.3% higher than we expect). Again, the shunt resistor with its 0.5% tolerance is likely to be the culprit.

After correcting the results, we get a trimmed error of $\pm 0.03\%$, very comparable with the ground-referenced circuit. However, our concern with the INA821 circuit is its performance over temperature.

The measured CMRR of this circuit was 106dB – not as good as the estimates of 120dB, but nevertheless acceptable. It's actually a bit difficult to measure CMRR, since things like op amp input offset voltage can also change over the common-mode range, and it's impossible to isolate the causes with a simple output voltage measurement.

The LT1167A circuit has worse un-

trimmed accuracy, peaking at almost 0.47%, but again the graphs show it to be almost all gain error. After trimming, the error is reduced to $\pm 0.025\%$, very similar to the INA821. The temperature coefficient is better, of course.

Another solution

Rather than commit to a \$30 chip, I want to introduce another trick we can use to improve precision in this type of situation.

So far, we have applied fixed offset and gain corrections to minimise the static errors in the circuit. In practice, this would be done for each sample in software, based on some one-off calibration performed at a standard temperature when we initially set up the instrument (and maybe when we periodically re-calibrated it).

Another approach might be to try to obtain the corrections in real-time at the ambient operating temperature. High-end instruments, like the 6½ digit multimeters that I used to obtain the

Table 5: theoretical improvement to Table 1 with dynamic zero trim		At Nominal 25°C		Additional error over 0-50°C (Nominal $\pm 25^\circ C$)	
Error	Nominal Value	Abs. Error	Rel. Error	Abs. Error	Rel. Error
Shunt Resistor: RES1 PCSR2512 (0.5%, 15ppm/°C)	100m Ω		0.50%		0.038%
Differential Voltage due to $I \times R_{SHUNT}$	100mV	0.5mV	0.50%	0.0375mV	0.038%
InAmp: INA821 ($V_{OS} \pm 35\mu V$, 5 $\mu V/^\circ C$) – zero trimmed	0mV	0.035mV		0mV	
InAmp Input Voltage total (Line 2 + Line 3)	100mV	0.535mV	0.54%	0.0375mV	0.038%
InAmpGain Resistor R_G : RN73C2A (0.1%, 10ppm/°C)	2k Ω	2 Ω	0.10%	0.5 Ω	0.025%
InAmp Gain Error (0.015% $\pm 35ppm/^\circ C$)			0.02%		0.088%
InAmp Gain (Line 5 $\times$ Line 6)	25.7	0.0296	0.12%	0.0289	0.113%
$V_{OUT DM}$ (Line 4 $\times$ Line 7)	2.57V	0.0167V	0.65%	0.0039V	0.150%
$V_{OUT CM}$ (20V, 120db, $\pm 1.5db$ over 0-50°C)	0V	0.02mV		0.0038mV	
V_{OUT} (Line 8 + Line 9)	2.57V	0.0167V	0.65%	0.0039V	0.150%

Difference & instrumentation amps

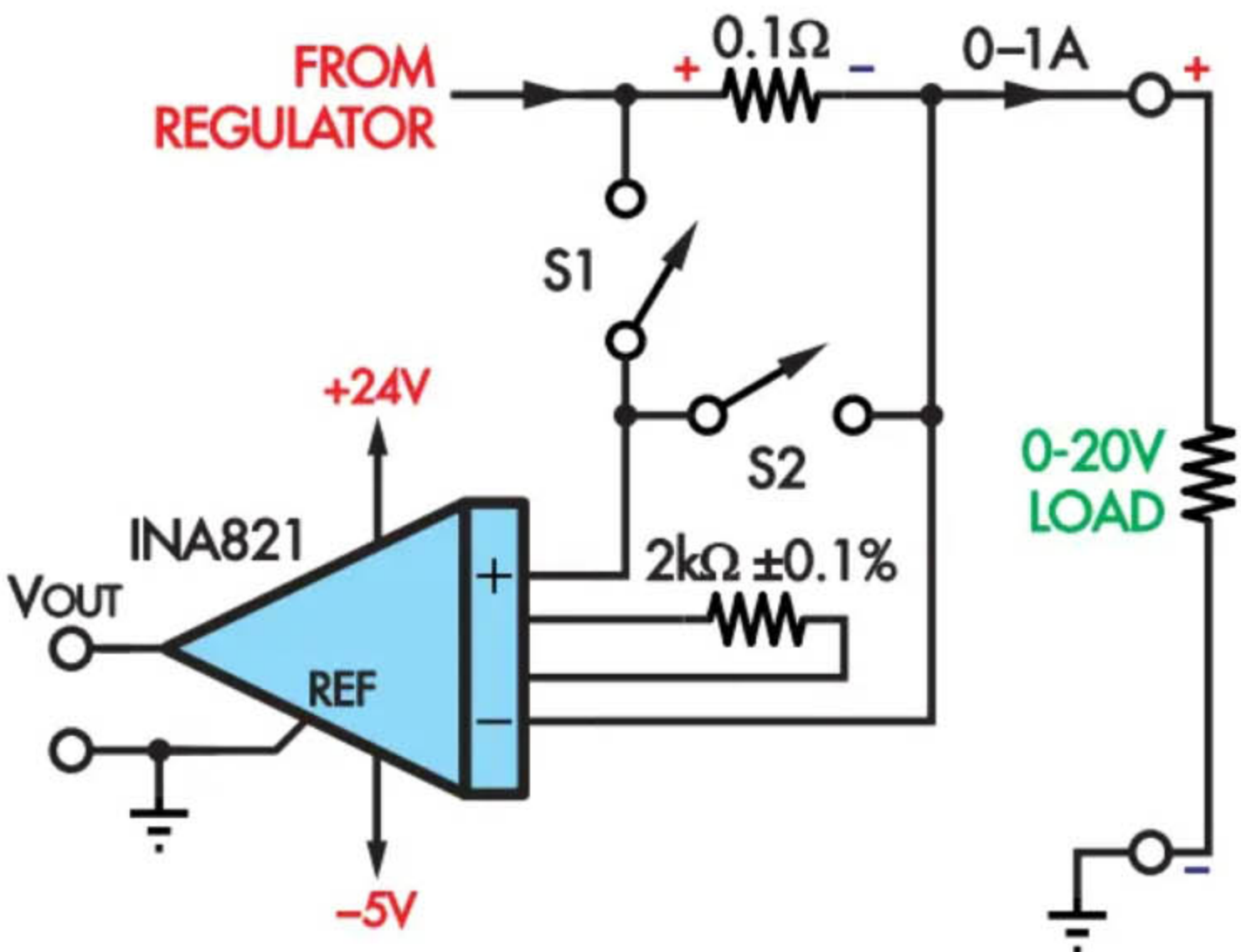


Fig.12: we can improve the temperature-dependent error of the circuit by adding switches to dynamically measure the offset, like an auto-zero op amp.

results shown here, effectively perform a zero and full-scale calibration every 20ms measurement cycle. Any temperature drift is calibrated out more-or-less in real-time.

We are not aiming for anything near that level of precision, but a simpler version can be a useful technique. It is pretty difficult to do a full-scale calibration of our test circuit, as we would need a precision 1A current source, but we could do a zero calibration fairly easily.

This won't let us trim out gain drift due to temperature but would let us calibrate out temperature-dependent offset errors in real-time – a bit like auto-zero op amps do.

Let's take a look at this approach using the INA821 example. Looking at the error budget table, we can see in line 3 that we have a possible $\pm 125\mu\text{V}$ drift in offset voltage over the temperature range. If we could calibrate that out, as shown in Table 5, we would almost halve the temperature error from $\pm 0.275\%$ to $\pm 0.15\%$.

Fig.12 shows one way we could achieve this in practice. Normally, S1 would be closed and S2 open so that we could take current measurements as before. Opening S1 and closing S2 shorts the inputs of the instrumentation amplifier so that we can use the ADC to read the circuit's offset voltage.

We would still need a fixed gain correction as before, but we can use the zero-scale reading to create a dynamic offset correction that will eliminate some of the temperature drift error.

Extending the range

Let's regroup and consider what we have achieved so far.

Measured Data		Untrimmed Error		Trimmed Error	
I (mA)	V _{OUT} (mV)	Absolute (mV)	Relative	Absolute (mV)	Relative
0.000	0.233	0.23	0.01%	0.48	0.019%
99.726	256.654	0.36	0.01%	-0.19	-0.007%
199.824	514.948	1.40	0.05%	0.05	0.002%
299.980	772.739	1.79	0.07%	-0.36	-0.014%
400.008	1031.164	3.14	0.12%	0.19	0.008%
499.980	1289.040	4.09	0.16%	0.34	0.013%
600.007	1546.980	4.96	0.19%	0.41	0.016%
699.965	1804.750	5.84	0.23%	0.49	0.019%
800.024	2062.770	6.71	0.26%	0.56	0.022%
899.971	2320.490	7.56	0.29%	0.61	0.024%
999.866	2578.110	8.45	0.33%	0.70	0.027%

Table 3 – untrimmed measured results from the INA821 circuit shown in Fig.8.

Measured Data		Untrimmed Error		Trimmed Error	
I (mA)	V _{OUT} (mV)	Absolute (mV)	Relative	Absolute (mV)	Relative
0.000	1.614	1.61	0.06%	0.46	0.018%
99.759	258.745	2.37	0.09%	0.14	0.005%
199.898	517.182	3.44	0.13%	0.14	0.005%
299.829	775.181	4.62	0.18%	0.24	0.009%
400.044	1033.716	5.60	0.22%	0.15	0.006%
500.013	1291.840	6.81	0.26%	0.28	0.011%
600.390	1549.970	6.97	0.27%	-0.64	-0.025%
700.009	1807.980	8.96	0.35%	0.28	0.011%
800.060	2066.110	9.96	0.39%	0.20	0.008%
899.975	2323.980	11.04	0.43%	0.21	0.008%
999.872	2581.780	12.11	0.47%	0.20	0.008%

Table 4 – untrimmed measured results from the INA821 circuit shown in Fig.8 when replaced with an LT1167A.

We have shown that with the shunt in the positive supply, we can probably achieve a trimmed accuracy of around 0.03% at 25°C with an additional 0.15% error over the 0–50°C temperature range if we use the INA821 instrumentation amplifier and dynamic offset correction. Let's call this 0.2% of total error.

This suggests we will have an overall resolution of $\pm 2\text{mA}$ in our 1A current (ignoring ADC precision for now). That is not good enough to measure the microamp resolution we would like to achieve.

I hope it is clear by now that we are not going to get the required three orders of magnitude improvement in precision just by improving the signal conditioning circuit. Even if we could, we will run into ADC quantisation limits, which we will cover in a later article.

The current circuit needs an ADC with at least 10 effective bits of resolution – three more orders of magnitude would require over 33 bits of effective resolution, which is pushing the limits of what is possible!

There is another way. We could pretty easily scale the range of the circuit by using a different shunt resistor. For example, using a 10Ω resistor would give a range of 0 to 10mA with $\pm 20\mu\text{A}$ resolution, while a 1kΩ resistor would yield a range of 100μA full-scale with $\pm 200\text{nA}$ resolution.

That will require some additional circuitry to switch the ranges. This, and the dynamic offset zeroing, will require us to add some switching elements to our signal path, which will themselves introduce some imprecision. We will look more deeply into signal switching in the next instalment of this series.

WEIRD & WONDERFUL ARDUINO PROJECTS

Part 3: building a 14×10 tricolour LED array

Hello once again. I can't believe it's March already! It seems like every time I blink, another day slips by and every time I sneeze, a month zips past. But enough of my ramblings.

I'm currently quivering in anticipation of what is about to unfold. We are poised to take the strip of 144 tricolour light-emitting diodes (LEDs) that we introduced in our previous column and use it to build the 14×10 array that will form our retro games console's main display.

Even if you aren't planning on building one of these beauties yourself (I have no idea how you could resist the temptation), you can still follow along with the rest of us because we'll be learning all sorts of useful tips and tricks along the way.

When working with semiconductor devices like integrated circuits (ICs) and LEDs, it's important to do everything on an anti-static mat while also wearing an anti-static wristband. This was discussed in *excruciating* exhilarating detail in *PE*, January 2023.

Rows and columns

Almost immediately, we are faced

with a choice. Should we refer to our main LED display as being a 10×14 array (height $\times$ width) or as a 14×10 array (width $\times$ height)?

In some fields, particularly mathematics and computer science (and in programming contexts), arrays and matrices are often described by the number of rows and columns, which translates to height $\times$ width. However, when looking at an image or a screen, our visual intuition is to think of its horizontal span (width) before its vertical span (height) because we read left-to-right, top-to-bottom.

When we come to writing our programs, we're going to be describing things in terms of their (*x*, *y*) coordinates. The main thing is for us to be clear and consistent; we will employ the (*x*, *y*) or *W* $\times$ *H* notation henceforth.

Chop chop

To make our discussions easier, we will refer to each of our tricolour LEDs as 'pixels' (an abbreviation of "picture elements"). Let's start by taking a slightly closer look at our strip (Fig.1). You should see white arrows indicating the direction of data from the input of each pixel to its output, and from

that output to the next pixel's input.

There are three copper pads between adjacent pixels. One is for 5V, one for 0V, and one for data. There are also copper wires inside the strip that aren't visible from the outside. I've used red and blue lines to show the 5V and 0V rails on the strip in Fig.1, because this will facilitate our discussions later.

However, real strips don't have these lines. Instead, they show '+' and '-' characters, respectively.

I think of the strip as having an 'input end' (shown here) and an 'output end'.

For our initial experiments last month, we connected a 390Ω resistor (with orange, white and brown bands) in series with the data input as close to the strip as we could. This protects the first pixel in the chain from the Arduino's enthusiastic switching speeds that can cause overshoot and undershoot on this signal if we're not careful.

Originally, we employed red (5V), white (0V) and green (data) wires for the signals feeding the strip, matching the colour scheme of the connector already attached to the strip. I've moved to showing red (5V), black (0V) and yellow (data) wires in Fig.1 because that is what I'm using in my console implementation, so that is what we will see in the photos I'll be showing.

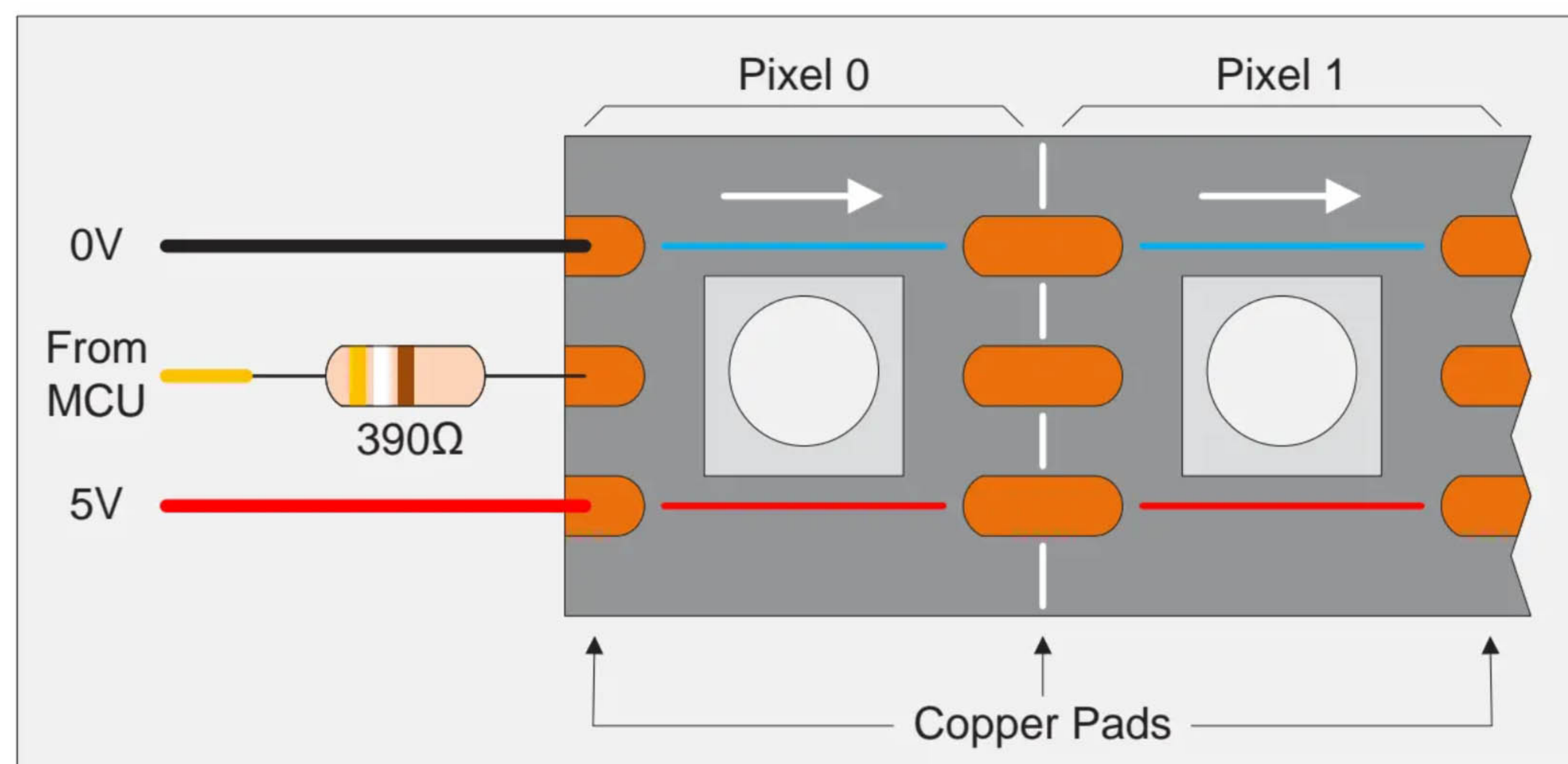
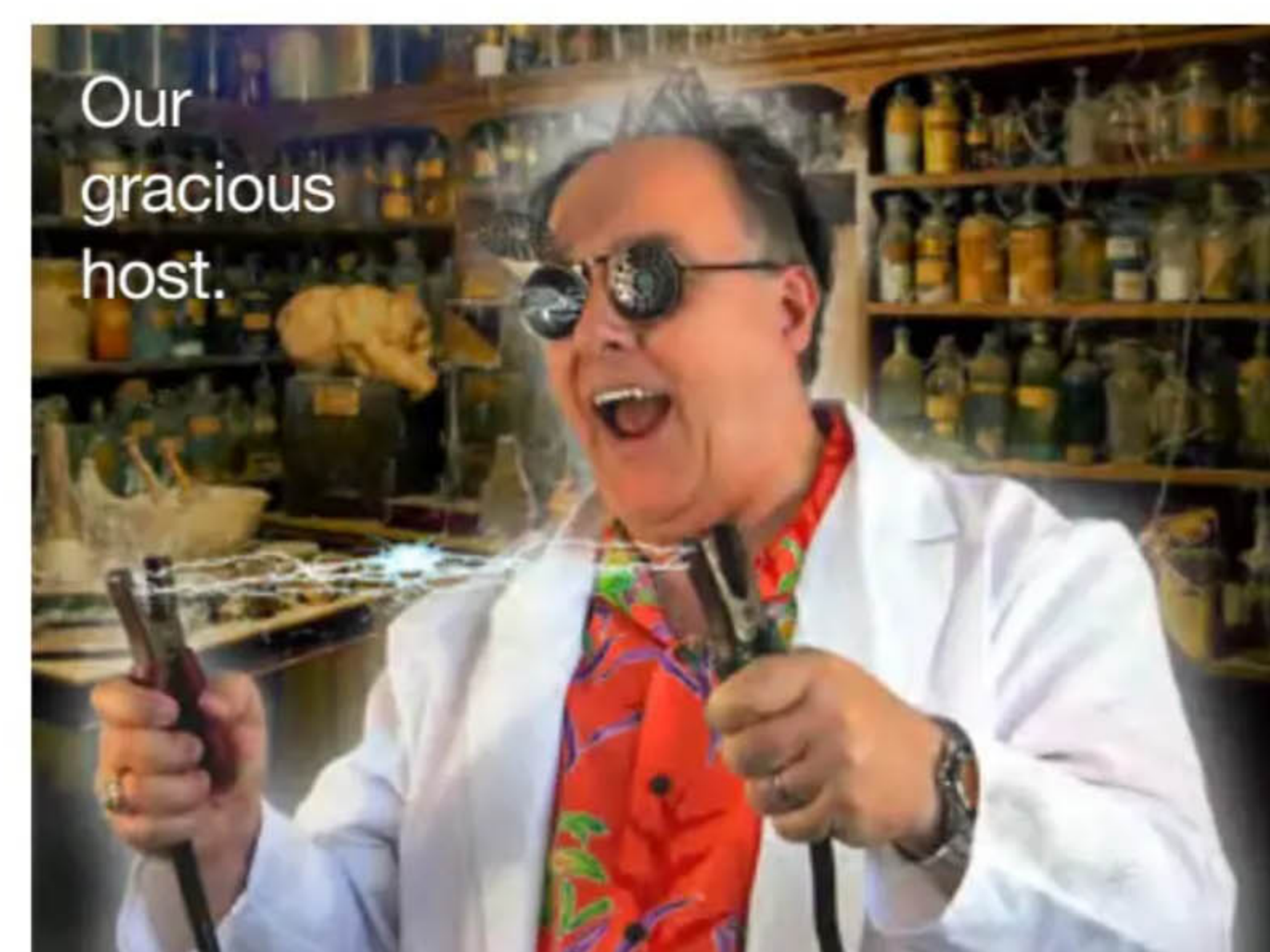


Fig.1: a close look at the strip.



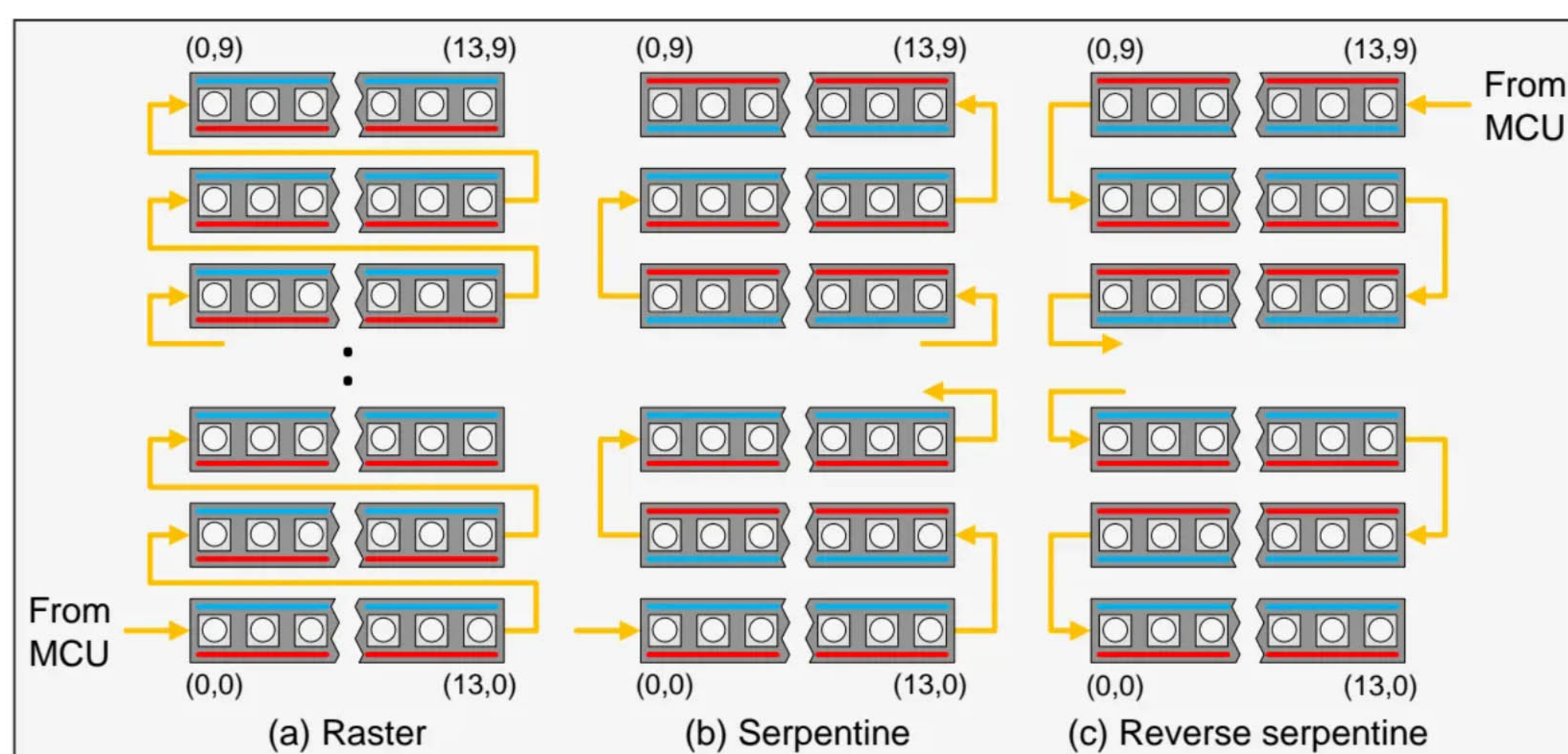


Fig.2: different ways of connecting our segments together.

Remember that our strip contains 144 pixels. We want to cut this into 10 segments, each containing 14 pixels. Even my poor maths informs me that this will leave us with four pixels left over.

Before you start cutting with gusto and abandon (and aplomb, of course), check your strip carefully, looking for a join somewhere in the middle. In my case, there was no join (hooray!).

I started cutting my 14-segment sections from the output end of my strip. Why? When I'd finished, this left me with a four-pixel strip that was still attached to my wires and resistor from last month. I can use this for future experiments.

What if you do have a join in the middle? This happened to my friend Joe Farr, who has already built a prototype of our console. Joe was lucky. His join occurred 44 pixels in from the output end of his strip, so he could cut three 14-pixel segments, then a 4-pixel chunk with two pixels on either side of the join, then continue to cut the remaining segments.

Things may be a little trickier if you have a join that prevents you from obtaining 10 'clean' 14-pixel segments. Is this really a problem? Possibly not. It depends on how you intend to construct your console.

It may be you are planning on mounting everything onto something like a sheet of hardboard or plastic. I thought that a cheap-and-cheerful clipboard could do the job (<https://pemag.au/link/ac3i>). In this case, the fact that a couple of your pixels are a little closer together—or, more likely, a little farther apart—won't matter all that much in the grand scheme of things.

On the other hand, if you are going for a full Monty implementation with a laser-cut front panel, you will need all your pixels to be evenly spaced and precisely aligned. In this case, you have only two options if your strip has an awkward join: (1) desol-

der, trim and re-solder the join until things are as close as you can get or (2) buy a second strip and use this to source the missing segment.

I used my trusty side cutters for the actual cutting. This may be a good time to note that I recently penned a three-column mini-series titled *The Beginner's Guide to Electronics: Tools—What to Buy First*, and I covered cutters in Part 2 (<https://pemag.au/link/ac3j>).

The copper pads are very small, so you need to be careful to cut them as close to the middle as possible. You'll also find that solder attaches better to these pads if you first deoxidise them by gently abrading (scuffing) them with fine-grit sandpaper.

My seemingly cunning plan

Let's assume that we have our ten segments sitting in front of us. The next question is how we are going to connect these together. Three possibilities (out of many options) are illustrated in Fig.2.

Remember that our 140 pixels are numbered from 0 to 139. At a low level in our program, pixel 0 will always be the one being driven by the Arduino, with the numbers increasing sequentially based on how the segments are joined.

By comparison, from our perspective, looking at the front of our console, we are going to view our array in terms of (x, y) coordinates. We are going to employ a 'bottom-left' origin system, in which $(x, y) = (0, 0)$ is in the bottom-left corner. This is standard in mathematics and Cartesian plane graphing, where positive x values extend to the right and positive y values extend upward.

We should probably note that computer graphics, screen displays and digital imaging applications often use a 'top-left' origin system. In this case, $(x, y) = (0, 0)$ would be in the top-left corner, x would increase to the right as before, while y would increase downward.

When Joe wired his array, he opted

to use what we might call a "raster" scheme, as shown in Fig.2(a). One advantage of this scheme is that all the segments are orientated in the same way, so the power and ground connections are in the same place for each segment. One disadvantage is the longer data signal wires required to connect the segments.

An alternative would be to use the "serpentine" scheme shown in Fig.2(b). The advantage of this scheme is the shorter data signal wires required to connect the segments. The disadvantage is that adjacent segments are reversed compared to each other. This is illustrated by the alternating positions of the red (5V) and blue (0V) lines on the strips.

I dare to be different. My plan was so cunning that we could have pinned a tail on it and called it a weasel, as Black Adder might say. I decided to use the "reverse serpentine" scheme illustrated in Fig.2(c). In this case, unlike the other two schemes, pixel 0 is at $(x, y) = (13, 9)$ rather than $(0, 0)$.

This turned out to be a not-so-cunning plan, more like something Baldrick would come up with. The only advantage of this scheme (and I'm clutching at straws here) is that it will provide an important lesson in a moment.

In many respects, it really doesn't matter how we decide to connect our segments, so long as we follow a regular repeating pattern. We can use our software to sort out the mapping of virtual (x, y) pixels to the physical pixels in our string later.

Attaching the segments

The strip of tricoloured LEDs I purchased came equipped with double-sided tape on the back. I'm assuming your strip is similarly endowed. If not, you'll need to attach some double-sided tape yourself, or you might opt for another attachment method like hot-melt glue.

As we previously discussed, you may decide to attach your segments—along with everything else comprising the console—to a sheet of hardboard or plastic. Alternatively, you may opt to use a laser-cut front panel and a 3D-printed case, or you might choose to do something in between.

I'll explain what I did; it's up to you to decide whether to follow my lead or to go your own way.

I was lucky because, when Joe was creating his prototype, he fabricated some extra bits and pieces, which he kindly donated to the cause (that would be me).

First, Joe provided me with a white laser-cut panel as seen in the centre

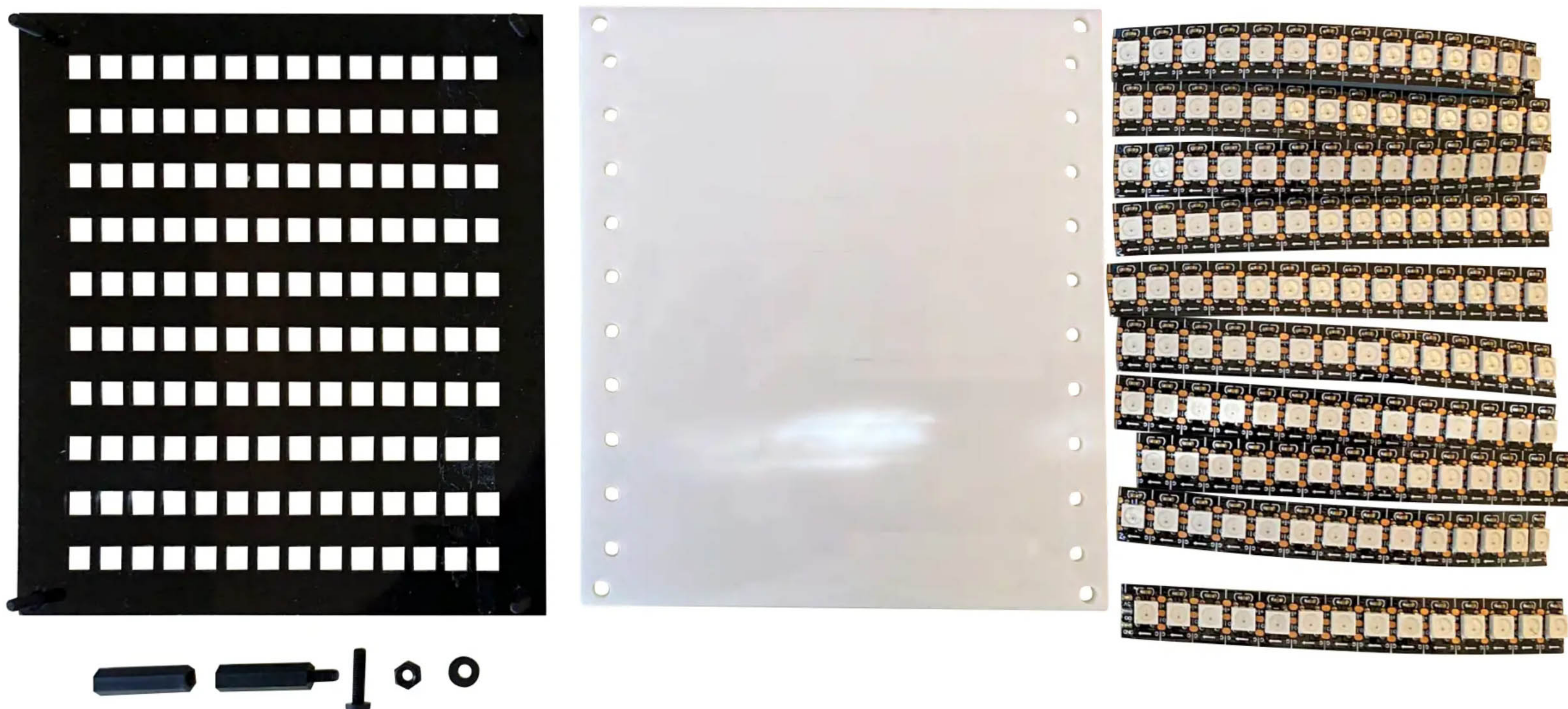


Photo 1: the inner fascia (left), mounting panel (center), NeoPixel segments (right) and joining hardware (bottom).

of Photo 1. This is the base to which I'm going to attach my segments. Observe the four mounting holes in the corners. In the fullness of time, we will use these to attach this board to the main fascia (front panel).

Next, observe the small black laser-cut 'inner fascia' panel shown on the left in Photo 1. We'll discuss the actual purpose of this panel later. For the moment, we're going to use it to align our segments, which you can see on the right in Photo 1.

Below the inner fascia in Photo 1, we see a few sample pieces from a printed circuit board (PCB) spacer kit. I just found a nice kit on Amazon (<https://pemag.au/link/ac3k>). This is an M3 kit, which means the nuts, bolts (machine screws) and pillars are metric with a nominal diameter of 3mm. M2.5 kits are also available. I usually opt for M3, but it's your call what you use.

The reason I mention these spacers here is that, if you look closely at the photo of the inner fascia, you can see that I've inserted four machine screws in the four mounting holes in the corners. In the not-so-distant future, we're going to use these screws to align the white board with the segments.

Next, we need to lay out our segments to match the ordering shown in Fig.2(c) (or whichever scheme you choose to use for your console), after which we flip them over and lay them on top of the inner fascia. Do this such that the pixels on the strips slot into the holes in the inner fascia. This leaves everything looking as shown in Photo 2.

Finally, we remove the blue backing from the double-sided tape, after which we drop the white panel over the four machine screws in the corners and press down gently, but firmly, to ensure the segments attach to the panel.

Segment data wiring

When we remove the white board, all the segments should be attached, but it won't hurt to run a finger along each segment pressing gently down on each pixel to ensure good, tight contacts.

The next step is to wire everything up as illustrated in Photo 3. Once again, your wiring will depend on the scheme you choose to use from Fig.2 (or some other arrangement of your own devising). I started with the green wires connecting the data signals between segments.

I could have used regular multi-core wire for this, but the green wires in a single-core wire jumper kit were the perfect size. I just found a similar kit on Amazon (<https://pemag.au/link/ac3l>). The colours of the wire lengths in that kit may be different to those in mine.

One point worth noting at this stage is that the bulk of the wiring takes place behind the board. Joe had located the holes for the wiring centred on his segments, which worked well for his connection scheme, as per Fig.2(a).

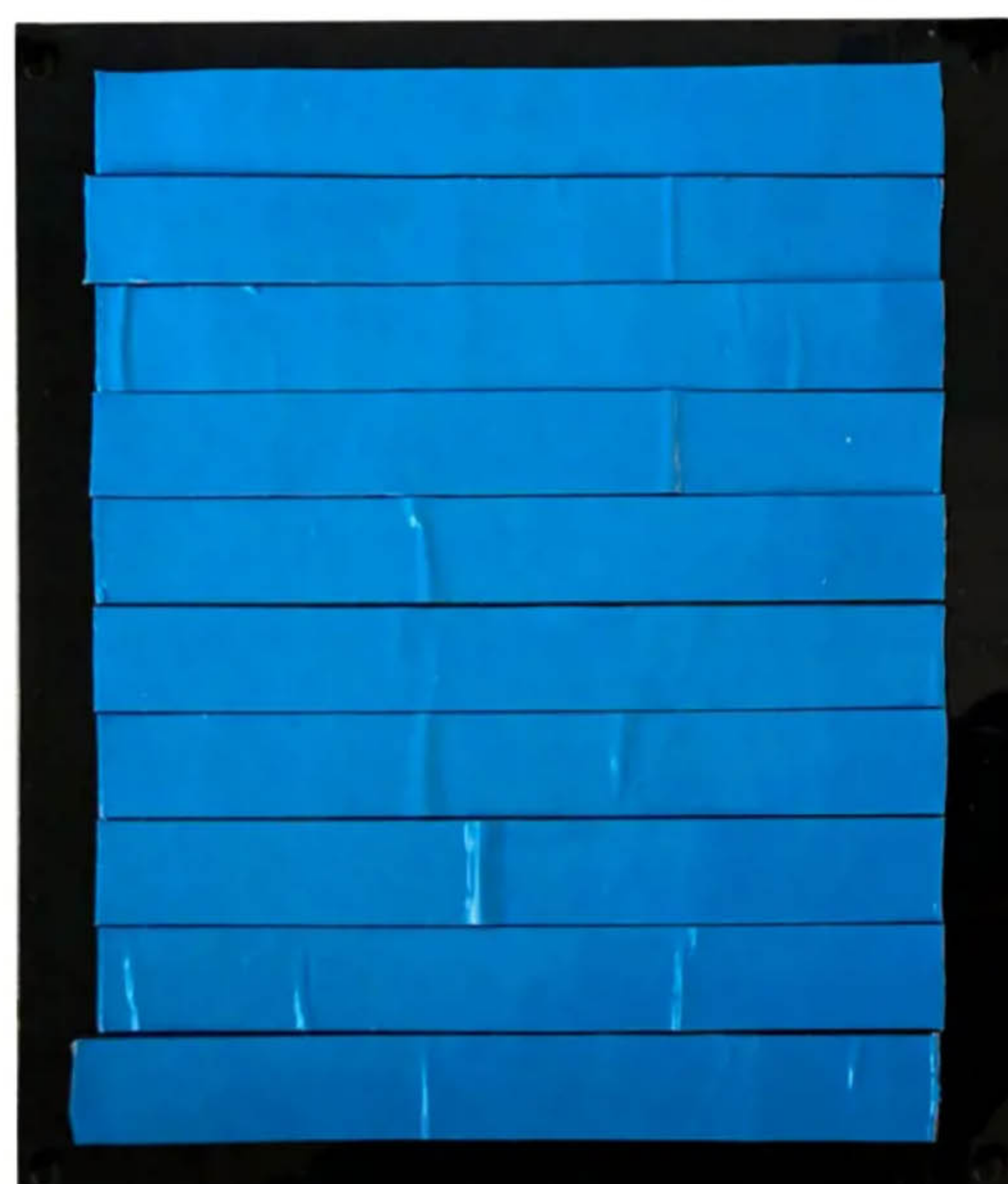


Photo 2: using the inner fascia to align the segments.

In the case of my not-so-cunning connection arrangement per Fig.2(c), my green solid jumper wires obstructed these holes. I would have been better off with holes between the segments. I could have easily added these myself, but I didn't realise there was a problem until after I'd soldered my green wires (rats!).

Segment power wiring

Next, we need to connect our red (5V) and black (0V) wires. What gauge (diameter) wires should we employ for this? This depends on how much current we intend to use. The data sheet states that each of the red, green and blue LEDs in each pixel has a maximum forward current of 20 milliamperes (20mA), giving a maximum possible current draw of 60mA per pixel.

If we were to drive all our pixels full-on, this would equate to $140 \times 60\text{mA} = 8400\text{mA} = 8.4\text{A}$. Eeek!

We will delve deeper into this topic in next month's column. For the moment, let's just say that we are

going to limit our current draw to a maximum of 10mA per pixel. Based on this, we know that each of our segments will draw only $14 \times 10\text{mA} = 140\text{mA}$.

We are going to use multi-core wire for this task because it's nice and flexible. A quick Google search reveals that 24 AWG (0.5mm diameter copper) wire can carry 1A, 22 AWG (0.65mm diameter copper) can carry 1.5A and 20 AWG (0.8mm diameter copper) can carry 2.5A.

In the AWG (American Wire Gauge) system, smaller numbers represent larger diameters due to the historical method used for wire manufacturing. The gauge number reflects how many turns of wire will fit on a rod of a certain length.

I opted to use 22 AWG (0.65mm) wire to attach to the segments themselves, but 24 AWG (0.5mm) would also have been OK. If you aren't already in possession of suitable wires, you could do worse than getting a kit containing a selection of different colours, like the one I just found on Amazon (<https://pemag.au/link/ac3m>).

Since our segments have only 14 pixels, we don't need to concern ourselves with the voltage drop, so I decided it would make my life easier if I connected the black (0V) wires on the left and the red (5V) wires on the right. Don't forget that the relative positions of these wires will be flipped between adjacent strips if you opt for a connection scheme like those shown in Fig.2(b) & (c)!

Keeping everything tidy

Now, let's look at the back of the board, as seen in Photo 4. We know that the maximum current for each segment will be 140mA. This means the maximum current if all ten segments are fully on will be $10 \times 140\text{mA} = 1400\text{mA} = 1.4\text{A}$. On this basis, we could use our 22 AWG (0.65mm) wire with its 1.5A capacity to gather all our segment wires and connect them to the main power supply.

On the other hand, it's better to be safe than to be sorry, so I opted to use 20 AWG (0.8mm) wires (<https://pemag.au/link/ac3n>) with their 2.5A capacity for this purpose. These are the single red and black wires exiting Photo 4 on the right.

Observe that the soldered joints where

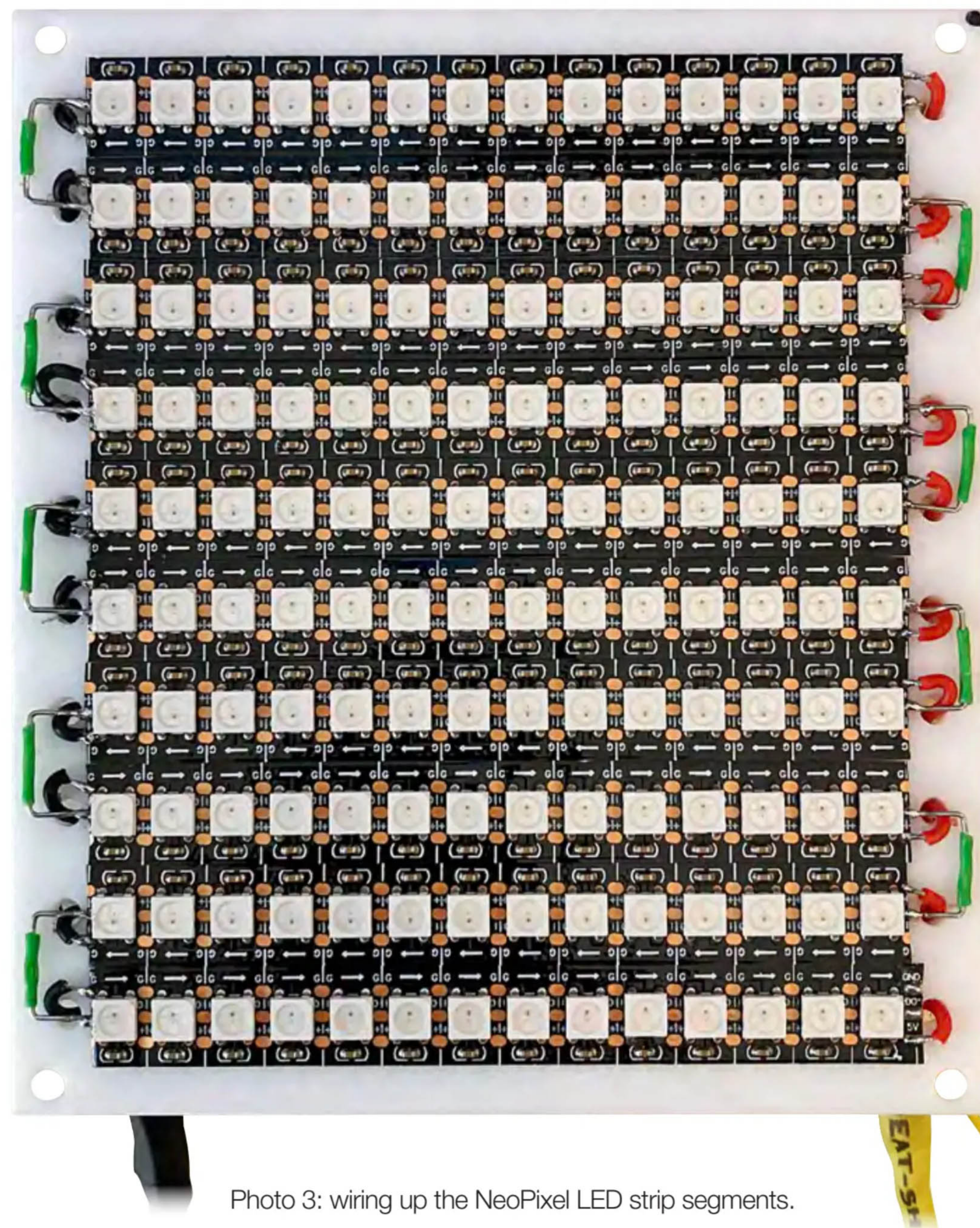


Photo 3: wiring up the NeoPixel LED strip segments.

these wires connect with the bunch of wires connected to the segments are covered and protected by heat-shrink tubing (we mentioned this last month).

The small piece of yellow heat-shrink tubing that appears in the bottom left-hand corner of Photo 4 is covering the 390Ω resistor that's connected to the data input of pixel 0 in my chain. The other side of this resistor connects to the 22 AWG (0.65mm) yellow wire we can see exiting from the bottom right-hand corner of Photo 4.

Since the yellow wire is carrying only a very low-current data signal, almost any gauge of wire would do, but 22 AWG is what I had on hand. Having more current carrying capacity than we need certainly doesn't hurt.

Rather than leaving my wires to their own devices, I created two wiring harnesses. That is, I bundled my wires together. I could have used cable ties but I decided to revive an old-cable lacing technique using waxed dental floss.

Besides making everything look nice and neat, these harnesses help to provide strain relief for the connection points where the wires are soldered to the segments.

Speaking of voltage drops

Earlier, I used the term 'voltage drop'. This refers to the reduction in electrical potential (voltage) as current flows through a component or a conductor (like a wire) with resistance.

The resistance of a wire is a function

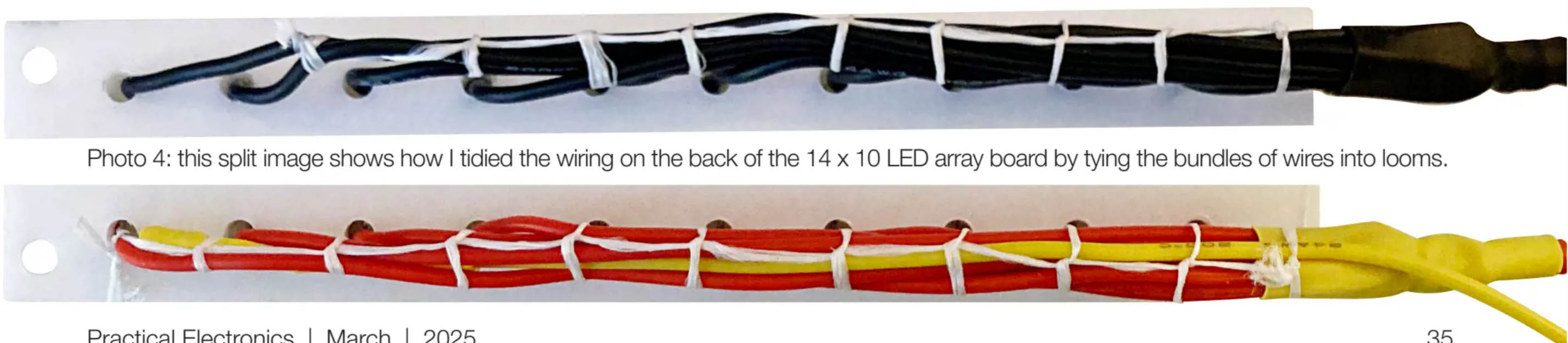


Photo 4: this split image shows how I tidied the wiring on the back of the 14 x 10 LED array board by tying the bundles of wires into looms.

of the resistivity of the material forming the wire and its length. The wires on our strip are formed from copper, which is an excellent conductor with low resistivity.

The resistance is directly proportional to the length of the wire (the longer the wire, the higher the resistance). It is also inversely proportional to the cross-sectional area of the wire (the greater the cross-sectional area, the lower the resistance).

The voltage drop from one end of a wire to the other is also a function of the amount of current flowing through that wire. To cut a long story short, (which is opposite to the way I usually like to do things), if we took a one-meter LED strip and all the LEDs were at 50% brightness, the voltage drop from the input end to the output end of the strip would be somewhere between 0.2V and 0.5V.

Alternatively, if all the LEDs were at 100% brightness, the voltage drop would be somewhere around 0.5–1.0V.

This means the LEDs would get dimmer as we progressed down the strip. For this reason, if we were powering a 1m strip, it would be a good idea to connect 0V and 5V wires to both ends.

Furthermore, if we were working with a longer strip—say 10 meters—we would adopt a ‘power injection’ approach. Specifically, we’d inject (connect) power (5V) and ground (0V) to the strip every metre or so.

In our case, since our segments are only 10cm long, and we are planning on running our pixels at a maximum of 10mA (which is only 16% of their maximum brightness), we don’t need to really be concerned with the voltage drop (phew!)

But does it work?

You can bet your booty it does! I took a copy of the second program that we used to test our strip last month. I changed the number of pixels from 144 (the full strip) to 140 (the number of pixels in our array) and dropped the RGB values from 255 to 32, which will give us white light at only 12.5% of full power (so we can see what it looks like).

You can download a copy of this code in the file named **CB-Mar25-Code-01.txt**. As usual, all the files mentioned in this column are available from the March 2025 page of the *PE* website: <https://pemag.au/link/ac3o>

As before, I connected the red and black wires from the array to the 5V and GND (0V) pins on the Arduino. Remember that we’re powering our Arduino via USB, which gives us 500mA. The Arduino itself requires 50mA, leaving us with 450mA to play with.

For the purposes of any experiments we will perform in this column, we will only ever activate a single row of 14 pixels at 12.5% of their full power, which means $14 \times (60\text{mA} \times 0.125) = 105\text{mA}$, so there are no problems there.

Next, I connected the yellow wire from the array to digital pin 12 on the Arduino, because this is the pin we’re using to drive the array in our program. Then I attached the USB cable to power everything up, downloaded my program into the Arduino and... held my breath.

Things occasionally don’t work when I power something up for the first time. I’m usually poised to kill the power, especially if I start to smell something iffy. Then I moan and groan and set to tracking down what I’ve done wrong. Now and then, however, the Fates smile upon me and it all works the first time. This was one such occasion. I still have a great big grin plastered across my face!

We need a function

The way I’ve connected my segments, as in Fig.2(c), means that the first pixel in my physical string, pixel 0, is in the top right corner of my array at $(x, y) = (13, 9)$. Meanwhile, the last pixel in my physical string, pixel 139, is in the bottom right corner of my array at $(x, y) = (13, 0)$. This is illustrated in Fig.3(a).

Because of this, our current test program starts by lighting the pixel in

the top right corner. Then the lit pixel races left and right as it works its way down the array, following the path (arrows) in Fig.3(a), until it conceptually falls off the end in the bottom-right corner.

Now, suppose I want you to light the pixel at $(x, y) = (4, 2)$. All we need to do is calculate that pixel’s number in the physical string. Off you go; I’ll wait here until you’ve finished.

Not that easy, is it? What we need is a function that accepts the x and y values of the desired location in the array as inputs and returns the number n of the pixel in the string as an output. I use the term “Neo” to refer to pixels in my programs because I always think of these as being Adafruit’s Neo-Pixels. On that basis, we’re going to call our function **GetNeoNum()** – see Fig.3(b).

The great thing about this is that it doesn’t matter if you organise your segments differently to mine. So long as we both have **GetNeoNum()** functions that work with our own arrays, and whose external interfaces are identical, all you will need to do to run one of my programs is to exchange my version of this function for yours.

Now for an algorithm

Sad to relate, the segment connection scheme I’m using as depicted in Fig.2(c) doesn’t make things as simple as one might hope. I’m a visual person, so the way I usually set about tackling a problem like this is to start by creating a handy-dandy diagram like the one shown in Fig.4.

In this case, the values in the boxes represent the pixel numbers in the physical string. Now that we are armed with our trusty diagram, all that remains is to use it to derive a cunning algorithm (somewhat above Baldrick’s level of cunning this time).

Let’s start with a few definitions. We might not use all of them, but I

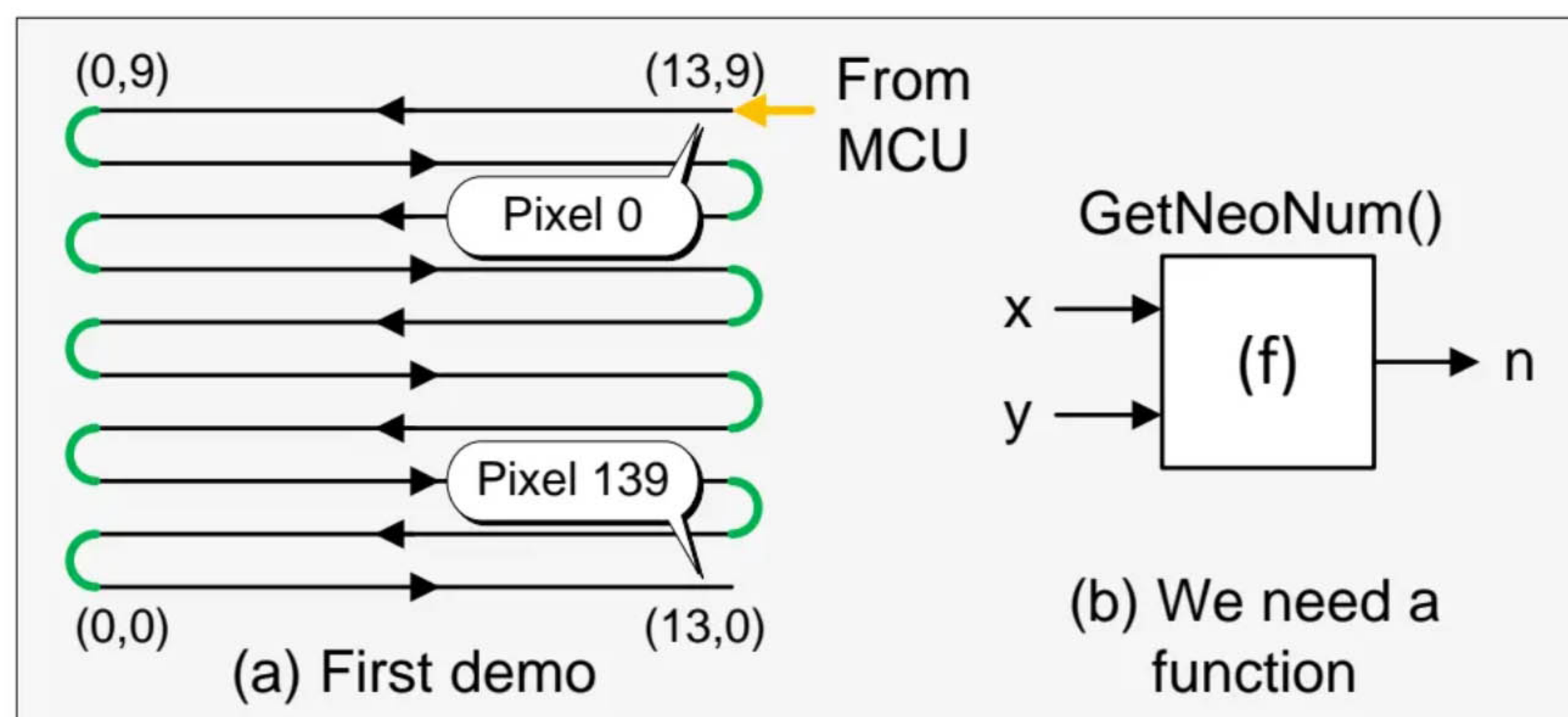


Fig.3: we need a function to convert (x, y) into a pixel number, n .

Columns														Rows
9	13	12	11	10	9	8	7	6	5	4	3	2	1	0
8	14	15	16	17	18	19	20	21	22	23	24	25	26	27
7	41	40	39	38	37	36	35	34	33	32	31	30	29	28
6	42	43	44	45	46	47	48	49	50	51	52	53	54	55
5	69	68	67	66	65	64	63	62	61	60	59	58	57	56
4	70	71	72	73	74	75	76	77	78	79	80	81	82	83
3	97	96	95	94	93	92	91	90	89	88	87	86	85	84
2	98	99	100	101	102	103	104	105	106	107	108	109	110	111
1	125	124	123	122	121	120	119	118	117	116	115	114	113	112
0	126	127	128	129	130	131	132	133	134	135	136	137	138	139
	0	1	2	3	4	5	6	7	8	9	10	11	12	13
	x													

Fig.4: this diagram shows the number for each pixel in my display.


```

50 // Return the number of the pixel in the string
51 int GetNeoNum (int x, int y)
52 {
53     int n;
54
55     if ( (y % 2) == 0 )
56     {
57         // Even Row
58         n = MAX_NEO - (y * NUM_COLS) - MAX_X + x;
59     }
60     else
61     {
62         // Odd Row
63         n = MAX_NEO - (y * NUM_COLS) - x;
64     }
65
66     return n;
67 }

```

Listing 2(a): the first iteration of my `GetNeoNum()` function.

always find this is a useful way to kick things off:

```

#define NUM_NEOS    140
#define MAX_NEO     139
#define NUM_ROWS    10
#define NUM_COLS    14
#define MAX_X       13
#define MAX_Y       9

```

As they say in America, this is not my first rodeo. Since my segments alternate back and forth, I know I'm going to need to treat the odd and even rows differently. Let's start with the lowest-hanging fruit in the form of row 0, column 0, which equates to $(x, y) = (0, 0)$. In this case, we want our function to return a value of 126, which is $139 - 13$. Hmm, that's $\text{MAX\_NEO} - \text{MAX\_X}$.

Now let's move to the next even row, which will be row 2. Row 2, column 0 equates to $(x, y) = (0, 2)$. In this case, we want our function to return a value of 98, which is $139 - (2 \times 14) - 13$. That's $\text{MAX\_NEO} - (y \times \text{NUM\_COLS}) - \text{MAX\_X}$.

Similarly, if we look at row 4, column 0, which equates to $(x, y) = (0, 4)$, we want our function to return a value of 70, which is $139 - 13 - (4 \times 14)$. Once again, that's $\text{MAX\_NEO} - (y \times \text{NUM\_COLS}) - \text{MAX\_X}$.

Now, let's return to row 0 but move to column 1, which equates to $(x, y) = (1, 0)$. In this case, we want our function to return a value of 127, which is $139 - 13 + 1$. That's $\text{MAX\_NEO} - \text{MAX\_X} + x$. In fact, for any of the columns in this row, all we need to do is subtract 13 from 139 and add our x value.

I think we're onto something here. We can generalise this to say that for even rows, ie, even values of y , our function should return $\text{MAX\_NEO} - (y \times \text{NUM\_COLS}) - \text{MAX\_X} + x$.

Performing a similar treatment for the odd rows, ie, odd values of y , our function should return $\text{MAX\_NEO} - (y \times \text{NUM\_COLS}) - x$.

Before we do anything else, let's take our algorithm for a spin. Do you remember earlier when I told you I wanted you to light the pixel at $(x, y) = (4, 2)$? Since this is an even row, we can say $n = \text{MAX\_NEO} - (y \times \text{NUM\_COLS}) - \text{MAX\_X} + x$. That equates to $139 - (2 \times 14) - 13 + 4 = 102$, which is what we'd expect from Fig.4. I'm performing my happy dance!

We can pick a few (x, y) values at random with different mixtures of odd and even values to verify that our algorithm works in all cases.

This is the point in the conversation where Joe would point out the advantages of his scheme from Fig.2a. The calculation inside Joe's version of the

`GetNeoNum()` function is easy peasy lemon squeezy: $n = (y \times \text{NUM\_COLS}) + x$ (seriously, that's all there is to it!). We will discuss all this in more detail next month, but for now we have other poisson à fire ("fish to fry").

Testing times

I just created a quick-and-dirty test program that employs our new algorithm; you can find it in the file named **CB-Mar25-Code-02.txt**. This features the first iteration of my `GetNeoNum()` function as illustrated in Listing 2(a). The listing number (2 in this case) corresponds to part of the name of the matching code file ("02" in this case).

Observe Line 55, where we use the $(y \% 2)$ modulus operation. This operator returns the remainder from an integral division. Since we are dividing by two in this case, the remainder can only ever be 0 or 1, and we can use this value to determine if y (the row) is even or odd.

For my first test, I decided to light all the pixels in each row in turn, starting with the bottom row and working my way up to the top row. It's **very important** to remember that we are currently powering our array from our Arduino (this will change next month), which means we only ever want to have a single row of pixels illuminated at a time.

As we previously discussed, this will draw 105mA. If we were to foolishly light all 10 rows at the same time, then—even though we're running each pixel at only 12.5% of its maximum capacity—this would equate to 1.05A, which would give our Arduino a 'bad hair day'.

The main `loop()` function in my test program as illustrated in Listing 2(b). In this case, the outer `for()` loop on Line 34 cycles through the rows. For each

```

30 void loop()
31 {
32     int n;
33
34     for (int y = 0; y < NUM_ROWS; y++)
35     {
36         for (int x = 0; x < NUM_COLS; x++)
37         {
38             n = GetNeoNum(x, y);
39             Neos.setPixelColor(n, Neos.Color(32, 32, 32)); // White
40             Neos.show();
41             delay(ON_TIME);
42             Neos.setPixelColor(n, Neos.Color( 0,  0,  0)); // Off
43             Neos.show();
44             delay(GAP_TIME);
45         }
46     }
47 }

```

Listing 2(b): the main `loop()` function in my test program.

row, the inner **for()** loop on Line 36 lights all the pixels on that row, waits for a delay of **ON_TIME**, and then turns the pixels off again before proceeding to the next row.

Why don't you try to come up with a program that extends this first test? For example, after lighting each row in turn from bottom to top, we could reverse things and light each row in turn from top to bottom. Then we could light each column in turn from left to right, finally reversing it and lighting each column in turn from right to left.

I just created a suitable program, in the file named **CB-Mar25-Code-03.txt**. It looks **great** running on my array. I cannot wait for you to see it running on your array (send pictures!).

Next time

We have so much to look forward to that I can barely contain my excitement. In our next column, we will start by lighting the pixels in diagonal lines across the display, one line at a time, starting with the one-pixel 'line' in the bottom left corner at (x, y) = (0, 0) and ending in the top right corner at (x, y) = (13, 9).

Next, we will move to powering our array with something other than our Arduino; something that will let us light all the pixels at the same time if we want to.

I must acknowledge that the current implementation of my **GetNeoNum()** function is horribly inefficient, so we will examine ways we can improve this little scamp.

My head is already buzzing with ideas and possibilities, but it would be great if you could also start pondering ways in which we could make the software better.

Also in our next column, we will perform some additional breadboard-based experiments with the eight-segment LED we introduced in the January 2025 issue. These experiments will

Useful Bits and Pieces from Our Arduino Bootcamp series

Arduino Uno R3 microcontroller module	https://pemag.au/link/ac2g
Solderless breadboard	https://amzn.to/3O2L3e8
8-inch (20cm) jumper wires (male-to-male)	https://amzn.to/3O4hnxk
Long-tailed 0.1-inch (2.54mm) pitch header pins	https://pemag.au/link/ac2h
LEDs (assorted colours)	https://amzn.to/3E7VAQE
Resistors (assorted values)	https://amzn.to/3O4RvBt
Ceramic capacitors (assorted values)	https://pemag.au/link/ac2i
16V 100µF electrolytic capacitors	https://pemag.au/link/ac2j
Momentary pushbutton switches	https://amzn.to/3Tk7Q87
Kit of popular SN74LS00 chips	https://pemag.au/link/ac2k
74HC595 8-bit shift registers	https://pemag.au/link/ac1n
Other stuff	
Soldering guide	https://pemag.au/link/ac2d
Basic multimeter	https://pemag.au/link/ac2f

Components for Weird & Wonderful Projects, part 1

4-inch (10cm) jumper wires (optional)	https://pemag.au/link/ac2l
8-segment DIP red LED bar graph displays	https://pemag.au/link/ac2c

Components for Weird & Wonderful Projects, part 2

144 tricolour LED strip (required)	https://pemag.au/link/ac2s
Pushbuttons (assorted colours) (required)	https://pemag.au/link/ac2t
7-Segment display modules (recommended)	https://pemag.au/link/ac2u
2.1mm panel-mount barrel socket (optional)	https://pemag.au/link/ac2v
2.1mm inline barrel plug (optional)	https://pemag.au/link/ac2w
25-way D-sub panel-mount socket (optional)	https://pemag.au/link/ac2x
25-way D-sub PCB-mount plug* (optional)	https://pemag.au/link/ac2y
15-way D-sub panel-mount socket (optional)	https://pemag.au/link/ac2z
15-way D-sub PCB-mount plug^ (optional)	https://pemag.au/link/ac30

* one required for each game cartridge you build
^ one required for each auxiliary control panel you build

Components for Weird & Wonderful Projects, part 3

22 AWG multicore wire kit	https://pemag.au/link/ac3m
20 AWG multicore wire kit	https://pemag.au/link/ac3n

include driving this device with one of the SN74HC595 shift registers we used in our *Arduino Bootcamp* series (PE, December 2024).

So, over to you. If you have any

thoughts that you'd care to share on anything you've read here, please feel free to drop me an email at max@clivemaxfield.com. Until next time, have a good one!

PE



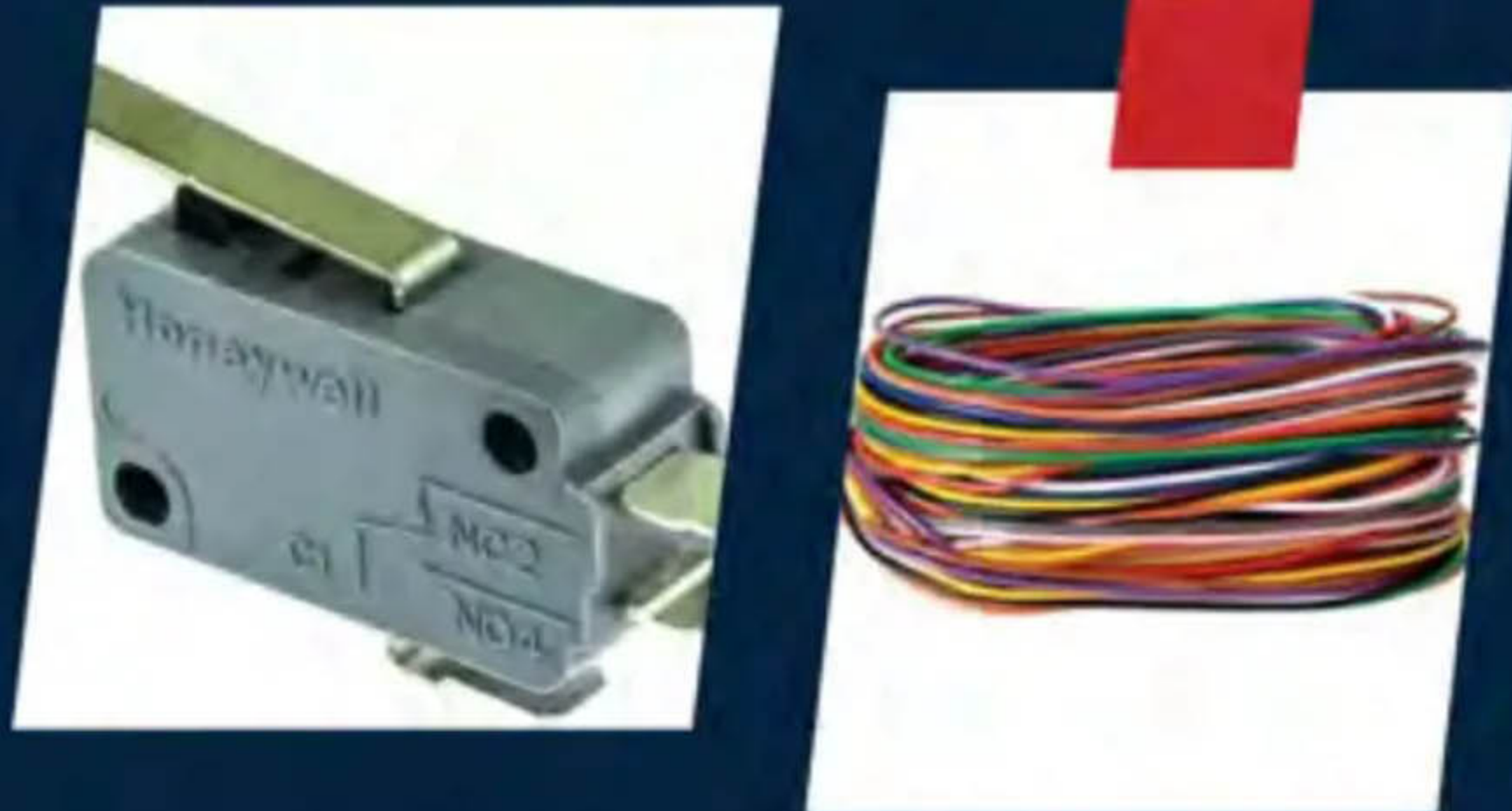
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Compact Frequency Divider

Project by Nicholas Vinen

This small board converts a standard 10MHz frequency reference (eg, from an oscilloscope) to 1MHz and 1Hz square wave signals. The latter can emulate the 1PPS output of a GPS receiver and has options for a 10% or 50% duty cycle.

Build your own compact Frequency Divider

10MHz – 1MHz | 10MHz – 1Hz



This straightforward circuit accurately divides a 10MHz signal to 1Hz with extremely low jitter. It has various applications, such as testing clocks and other devices that are time-locked to GPS signals.

It could also be used to drive several clocks from a single accurate time source or to derive a very accurate 1Hz signal from a low-cost 10MHz temperature-compensated crystal oscillator (TXCO).

The divider uses just four logic ICs, including the somewhat unusual 74HC4059, plus an ultra-high-speed comparator and a buffer for driving the outputs. It can be powered at 5V DC from a USB supply or 6.6-12V DC, drawing only about 10mA. It has supply reverse polarity protection and input overload protection and won't generate an output unless it's being actively fed a signal.

A jumper selects between the 10% and 50% duty cycle options for

the 1Hz output. The output jitter is extremely low as long as the input signal is relatively clean. Many pieces of test equipment will have a suitable 10MHz output, or you could use one of our GPS-Disciplined Oscillators:

- GPS-Disciplined Oscillator (May 2024)
- GPS-synched Frequency Reference (October and November 2019)
- GPS-based Frequency Reference (April & May 2009)

Circuit details

Its circuit is shown in Fig.1. We have tried to keep it simple and inexpensive without sacrificing performance. The 10MHz signal is fed into SMA connector CON1 and goes into the first stage, based on ultra-high-speed comparator IC6 (TLV3501).

The TLV3501 is an interesting device as it runs from 2.7 to 5.5V, drawing just 3.2mA and yet has extremely low input bias currents at $\pm 2\text{pA}$ (typical), a low input offset voltage of $\pm 1\text{mV}$

(typical) and very high-speed operation with a maximum toggle frequency of 80MHz. That makes it suitable for many applications.

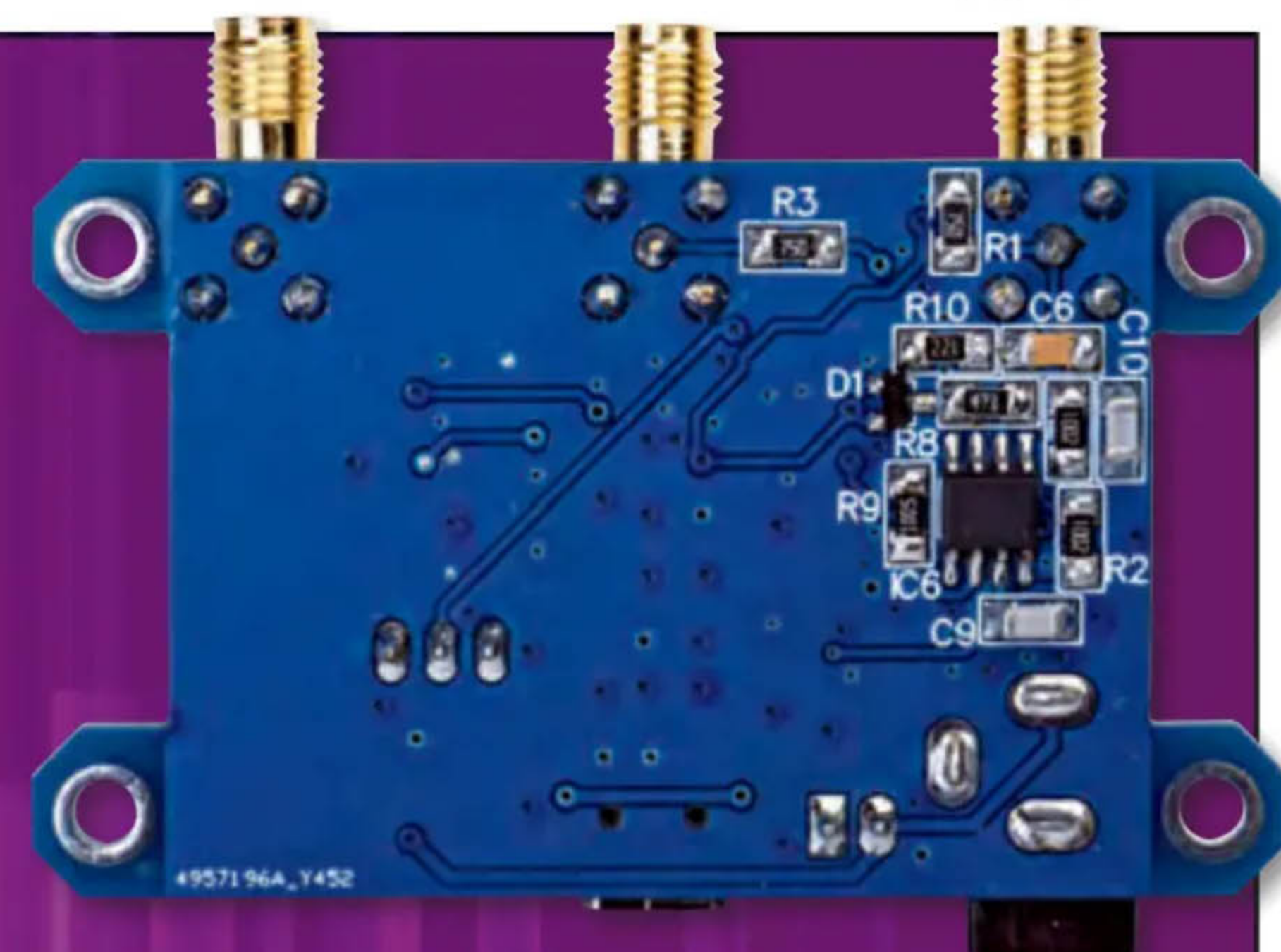
Here, its job is to convert what might be a relatively low-level, sinusoidal input signal into a 5V peak-to-peak square wave. That means the circuit is not too sensitive about what drives it, as long as it is a 10MHz waveform of at least 10mV RMS or 35mV peak-to-peak; it will most likely be a sine or square wave.

The circuit is designed with 75 Ω impedances in mind, although you could change that if necessary (eg, using 49.9 Ω or 51 Ω resistors instead of 75 Ω). So the input is terminated with a 75 Ω resistor, then coupled to comparator IC6 by a 1nF DC-blocking capacitor and 220 Ω series resistor.

Dual schottky diode D1 protects IC6 from over-voltage or having a signal applied while the circuit is powered down by clamping the input signal to within 0.3V of the supply rails.

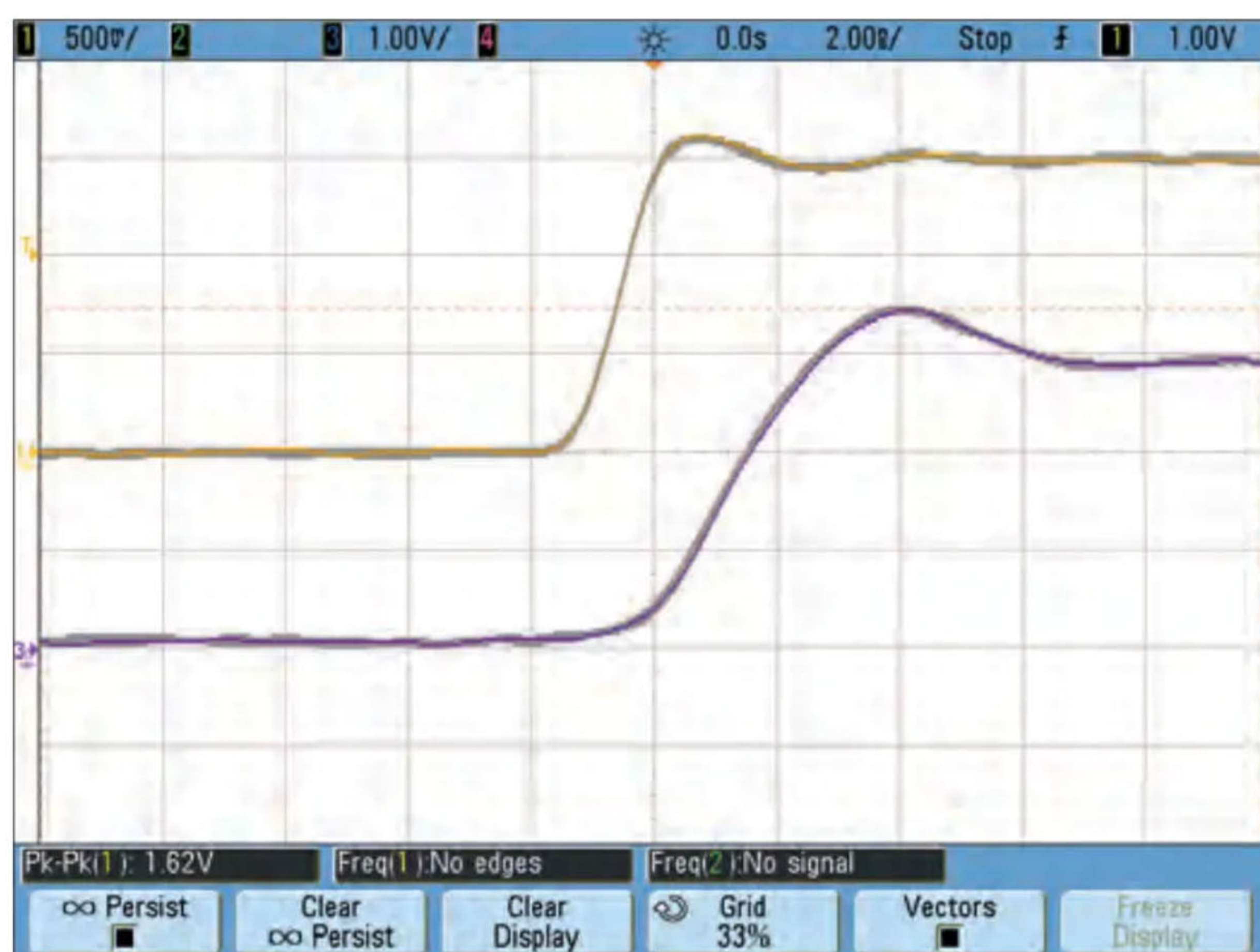
Frequency Divider Features & Specifications

- » Divides the nominally 10MHz input frequency by 10 (to 1MHz) and 10^7 (to 1Hz)
- » 10% or 50% duty cycle option for 1Hz output (5V peak-to-peak unloaded)
- » Operating input signal level: 10mV to 3.2V RMS (28mV to 9V peak-to-peak)
- » Recommended input signal: 35mV to 2V RMS (100mV to 5.6V peak-to-peak)
- » Jitter: estimated at 0.1ns with a clean clock source (see Scope 1)
- » Propagation delay: approximately 100ns
- » High noise immunity with 23.5mV built-in hysteresis
- » Outputs are in phase with inputs
- » No output signals if the input is not driven
- » SMA connectors for input and outputs
- » Choice of 50 Ω or 75 Ω input/output impedances
- » Power supply: 5-12V DC @ 10mA
- » Power connectors: USB Type-C, 2.1mm/2.5mm inner diameter barrel plug, (polarised) pin header
- » 3mm mounting holes: 4 (the board can be made smaller by cutting them off)

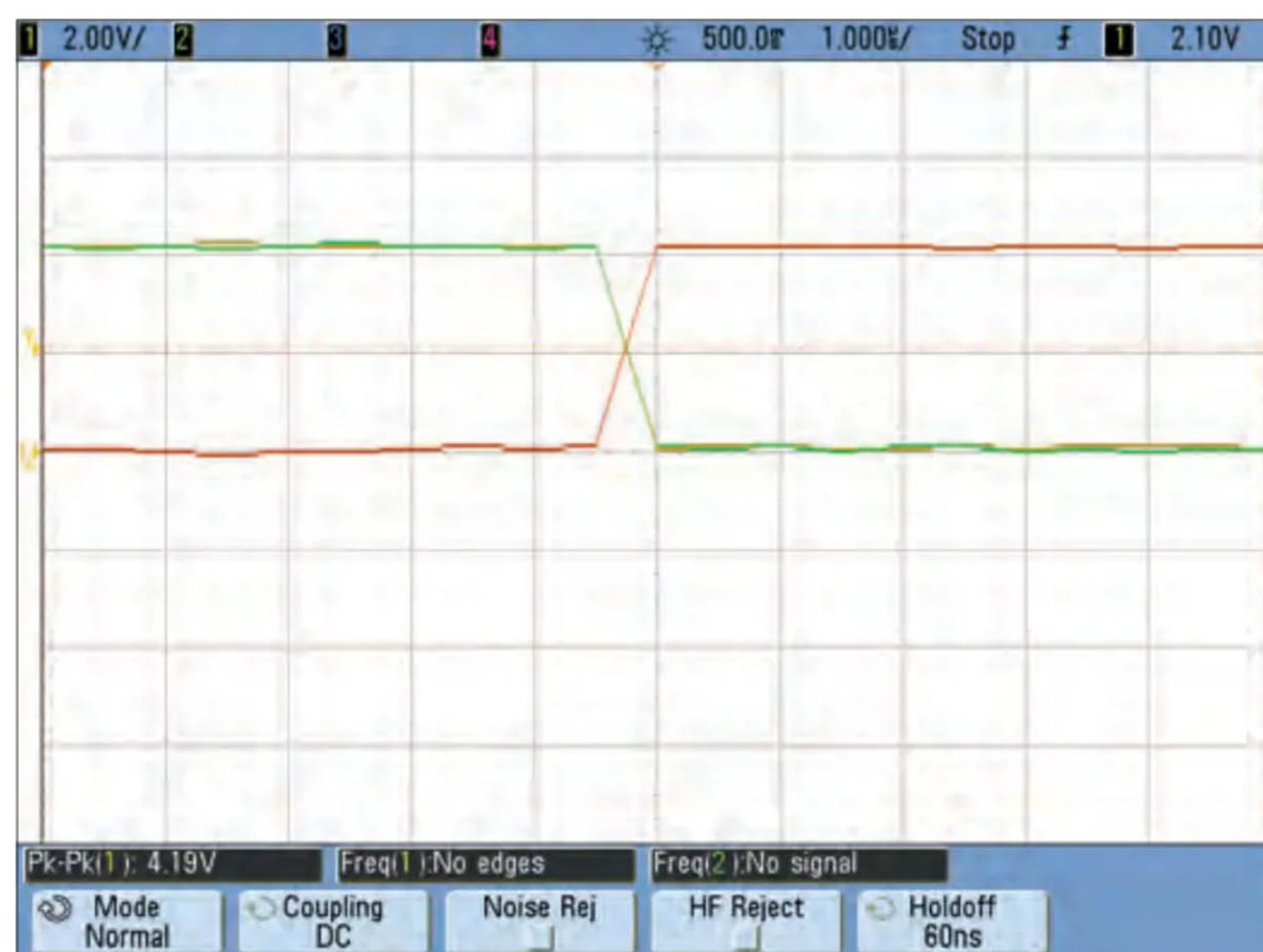


The prototype board is very similar to the final version; it just lacks the power LED and used a Type-B mini USB socket instead of the now more standard Type-C.

Constructional Project



Scope 1: the yellow waveform is the 1Hz output (reduced in amplitude due to a lower than normal supply voltage and 50Ω termination) while the blue waveform is the 10MHz reference signal from the oscilloscope. The grey areas around them show the previous 50 or so traces, indicating extremely low variation in timing between them (ie, low jitter). The output edge seems to come first because the propagation delay is just under the input signal period (100ns); it was triggered by the previous edge.



Scope 2: to check the frequency ratio was correct, we captured the unit's output on the scope for two seconds and then measured the time between edges. Here three of the captured edges are overlaid, in yellow (-500ms), red (0ms) and green (+500ms). The yellow and green traces overlap, indicating they are exactly one second apart as per the scope's timebase (and hence 10MHz reference oscillator). As we captured two full seconds of data, the time resolution is more coarse than in Scope 1.

The 220Ω resistor primarily exists to limit the current through these diodes, protecting them and the rest of the circuit from excessive 'bus pumping' of the 5V rail.

As the signal is AC-coupled to IC6, it is DC-biased to half of the 5V supply using a pair of 10kΩ resistors and a 47kΩ bias resistor. The 100nF capacitor prevents supply ripple from coupling back into the signal, which could cause jitter.

A 10MΩ resistor from output pin 6 of IC6 back to its non-inverting input, pin 3, provides around 23.5mV of hysteresis for noise rejection. This forms a voltage divider with the 47kΩ bias resistor for pin 3. With around 2.5V across the hysteresis resistor (regardless of whether IC6's output is at 5V or 0V), 2.5μV flows through it and subsequently the 47kΩ resistor, causing an offset of around 11.75mV.

That offset switches polarity as IC6's output switches, meaning that any noise on the input signal would have to exceed 23.5mV to cause an unwanted edge at IC6's output. It also means that there needs to be at least a 23.5mV peak-to-peak signal applied to IC6 before its output will start to toggle. Thus, it won't oscillate without a signal at CON1.

With the 75Ω termination resistor, plus the low-pass filter formed by the 220Ω resistor and IC6's input capacitance (plus that of both diodes in

D1), the minimum signal the circuit will respond to is about 30mV peak-to-peak at CON1. However, a higher level is recommended to ensure jitter-free operation. 30mV at CON1 implies a higher voltage at the signal source, probably closer to 60mV peak-to-peak.

Frequency divider

Now that we have a clean 10MHz square wave signal from IC6, it's fed to the first divider, IC1. This is a 74HC4017 Johnson decade counter, a lower-voltage, higher-speed version of the good old 4017 counter IC.

These are inexpensive, run from 2-6V, operate at up to 77MHz with a 5V supply and provide ten 10% duty cycle outputs with different phase angles, plus a single 50% duty cycle output that's phase-aligned with the input (and Q0 output).

For IC1, we feed the 10MHz signal into the pin 14 clock input and get a nice 1MHz square wave from the $\overline{Q5}$ - $\overline{Q9}$ output. The MR (master reset) line is tied low for constant operation, while the inverting clock input at pin 13 is also tied low as we are using the non-inverting clock input. The ten phase outputs, Q0-Q9, are not used in this case.

The 1MHz output from pin 12 is fed to two places: firstly, to three of the six buffers in IC5 connected in parallel, then to the 1MHz SMA output (CON2) via a 75Ω impedance-matching resistor.

The MC74VHCT50A is similar to a 74HC04 hex inverter IC except that it does not invert the signals but merely buffers them. That keeps the outputs in phase with the 10MHz input.

Secondly, the pin 12 1MHz output of IC1 goes to another 74HC4017 counter, IC2, configured identically to IC1. It produces a 100kHz square wave at its pin 12 output, which is fed to the clock (CP) input, pin 1, of IC3.

This is the 'main event', configured to divide its input frequency by a factor of 10,000. It is a larger IC than the others, with 24 pins rather than 16, and somewhat more expensive (but still pretty reasonable). It takes up less space than four more 74HC4017s and has a much lower propagation delay.

It can be configured for thousands of different frequency division ratios in various ways based on the logic states of its KA-KC and J1-J16 pins. The accompanying panel explains how this particular configuration achieves the 10,000:1 division ratio.

We could have added a microcontroller to this board, driving all those pins, and provided a few different ratios. However, we decided it was better to keep this simple and avoid programming any chips. Still, we will present a more complex programmable design that includes a microcontroller in an upcoming issue.

Compact Frequency Divider

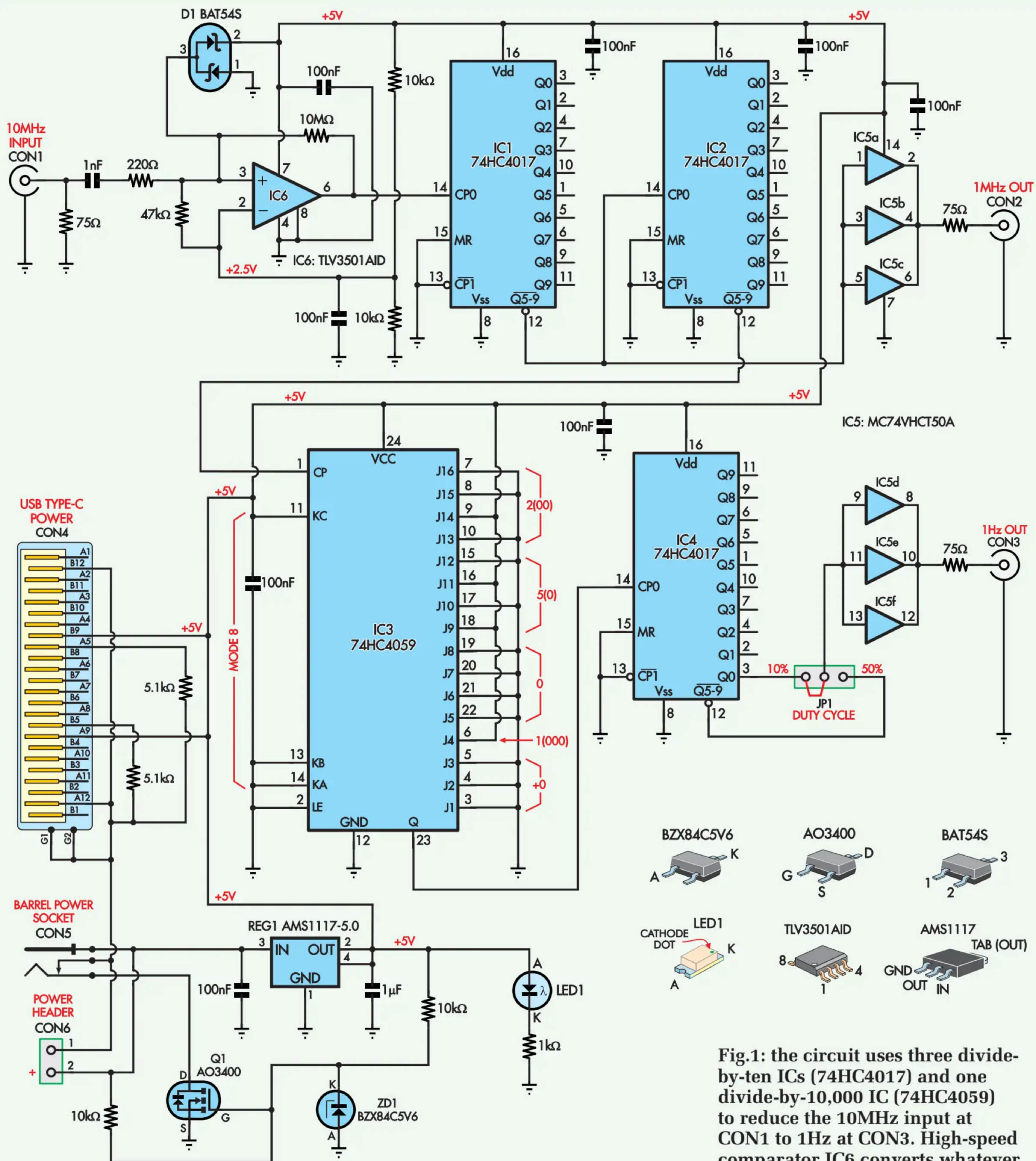


Fig.1: the circuit uses three divide-by-ten ICs (74HC4017) and one divide-by-10,000 IC (74HC4059) to reduce the 10MHz input at CON1 to 1Hz at CON3. High-speed comparator IC6 converts whatever waveform is fed in to a 5V peak-to-peak square wave for driving IC1.

IC3 has an output latch that we do not use, so the latch enable (LE) input, pin 2, is tied to ground. The 10Hz signal appears at pin 23 (Q). Note, though, that this pin will only be high for one input pulse, and with a 100kHz input, the output pulses are 10 μ s wide. That is why we divided the 10MHz signal by a

factor of 100 first; otherwise, the output pulses would be a mere 100ns wide.

To make this short pulse useful, we feed it to the final counter, IC4, another 74HC4017 configured much like the others. It performs the final division to get a 1Hz signal and converts the short pulses into a 50% duty cycle square

wave at its pin 12 output.

We feed that, plus the similar but shorter 10% duty cycle pulse from output Q0, to a three-way pin header. That allows you to select the desired duty cycle using a jumper shunt. The resulting signal is fed to another triple parallel buffer (IC5d-IC5f) and then the

Constructional Project

final SMA output, CON3, via another 75Ω impedance-matching resistor.

The 10% duty cycle output more closely simulates a GPS 1PPS output, while the 50% duty cycle signal is nice and symmetrical for driving something like a clock.

Power supply

There are three power supply inputs. The USB Type C connector (CON4) is the simplest as it feeds the USB 5V directly into the circuit. However, note that its ground connection goes via the internal switch in barrel socket CON5. This way, if you plug both in simultaneously, you won't be feeding power into the device connected to the USB socket.

Unlike USB Type-B sockets, the Type-C socket needs two 5.1kΩ pull-down resistors connected to signal the power source to deliver 5V. You can leave those resistors off the board if you aren't fitting the Type-C socket. This particular socket only has the six pins needed for USB power delivery, without the data signals.

By the way, we're switching from

Type-B to Type-C because it is now the universal standard, so expect to see more of this in future.

After passing through CON5's internal switch, the GND connection from CON4 also passes through Mosfet Q1 before reaching circuit ground. This provides reverse supply polarity protection, although that should not be necessary for the USB socket as the socket itself should guarantee the correct polarity. However, it is helpful if powering the circuit via barrel connector CON5 or header CON6.

In those cases, as Q1's gate is connected to the +5V rail and incoming DC supplies via two 10kΩ resistors, it will only conduct if the incoming supply polarity is positive. If it is negative, Q1's gate will be pulled negative, Q1 will be off, and the whole circuit will be unpowered, floating at the positive DC supply voltage (that was erroneously connected to the negative input).

There are two 10kΩ pull-up resistors for the gate so that Q1 will switch on regardless of whether the USB connector is used (feeding 5V directly) or one of the other inputs, which feed 5V

low-dropout regulator REG1. Zener diode ZD1 prevents damage to Q1 as its gate is only rated to handle ±12V.

This method has a much lower voltage loss than using a series diode (a few millivolts instead of 300mV+), allowing you to use a supply barely above 5V while still getting a regulated 5V at the output of REG1 to power the rest of the circuit.

Construction

While it uses mainly SMD parts, the board is relatively easy to assemble as they are all fairly large. Experienced constructors can gather the parts and solder them to the board as shown in overlay diagrams Figs.2 & 3. We suggest fitting all the SMD parts to one side of the board, followed by the other, then the through-hole parts. It's best to start with the top, as more parts are on that side.

The Frequency Divider is built on a double-sided PCB coded 04112231 that measures 64 × 37.5mm. We recommend soldering IC1-IC5 in numerical order, then ZD1, Q1, REG1, the USB socket (if fitting it), then the top-side capacitors and resistors.

With the ICs, check very carefully that each one is the right way around before soldering them; most will have a pin 1 dot or bar. Use a magnifier to find them if necessary. As shown in Fig.2 and on the PCB, in each case, pin 1 faces towards the top of the board or to the left (for IC5).

There is one 1μF capacitor on this side; the rest are 100nF. As mentioned earlier, you can leave off the 5.1kΩ resistors if you aren't using the USB socket. Also note that unlike the Type-B USB sockets we've been using for a while, these Type-C sockets have no locating posts that slot into holes in the PCB, so you will have to be careful to align all its pins and tabs with the pads before soldering more than one.

There are various ways to solder these parts: with solder paste and hot air, solder paste and a reflow oven, solder paste and a hot plate or regular solder and a regular iron (which is how we did it). If using a standard iron, we strongly recommend having a good quality flux paste on hand, plus some solder wick, as they make it much easier.

There is no 'right' way to hand solder SMD ICs, but here is how we did it, starting with the ICs. We placed a little solder on one of their pads, then

Programming the CD74HC4059 counter

This counter is quite complicated as it includes a prescaler plus a three or four digit 'decimal' main counter that varies in how you can use it. The prescaler can divide by between 1 and 10 in five different modes. However, which prescaler mode you choose affects what values you can have in the main counter's top (thousands) digit.

For example, if you have a divide-by-10 prescaler, the main counter only has three digits (up to 999). If you use one of the other prescaler values, the main counter has four digits, with more options as the prescaler division ratio becomes smaller.

The lower three decimal digits of the main counter can always be preset with a value from 0 to 9. Depending on the mode, the overall maximum division ratio is either 9999 (eg, with the prescaler in divide-by-10 mode) or 15999 (with the prescaler dividing by a power of two).

It is actually possible to divide by a much higher number than that because the 'BCD' or 'binary coded decimal' counter stages that it initially seems can only count up to 10 are actually full binary counters that can count up to 16. So, while programming it is trickier, it can be set to divide by up to 21,327.

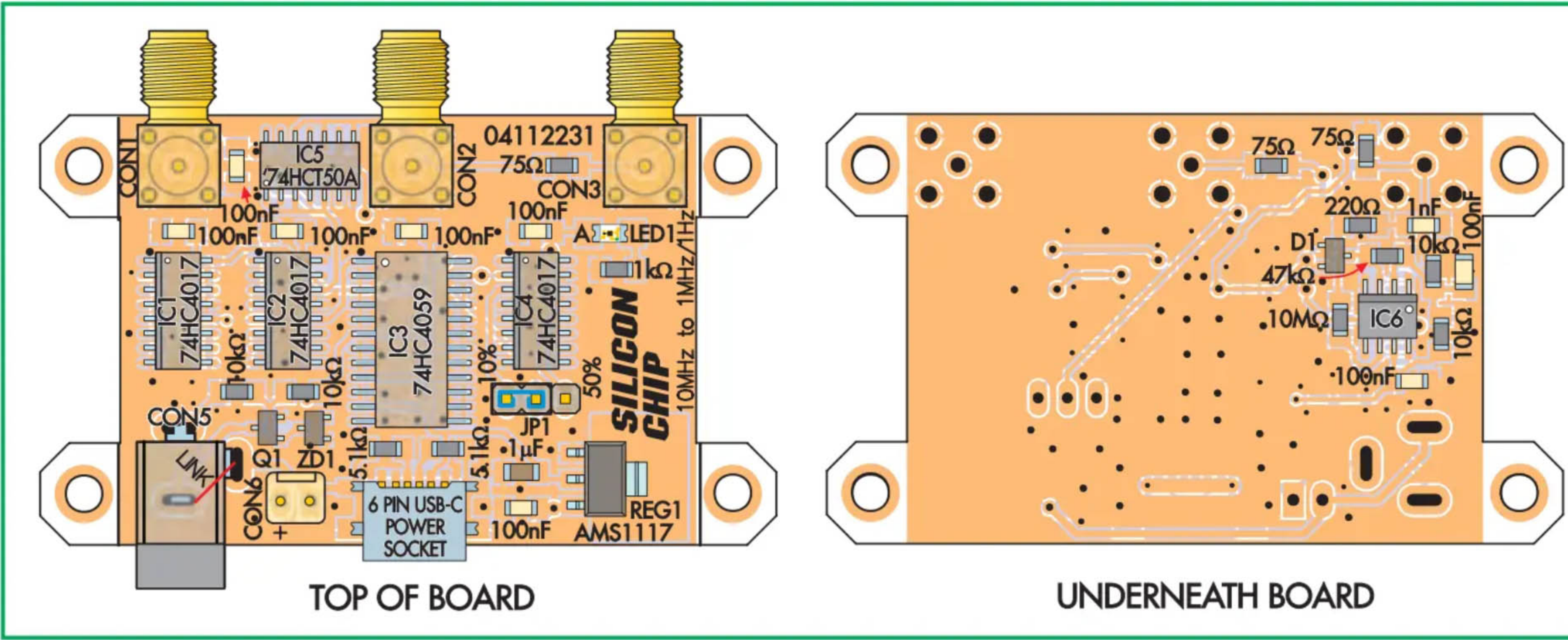
Luckily, our desired division ratio of 10,000 is relatively easy to set up. We could have used a prescaler value of 10, leaving a three-digit main counter. While dividing by 1000 with three digits seems impossible, we could have set the top 'digit' to 10 (because the actual limit is 15), which would have given the desired result.

In the final design, we use a prescaler ratio of 8, leaving us with four digits for our main counter, although the top digit can only be 0 or 1. That's fine because we set the main counter to divide by 1250, as $1250 \times 8 = 10,000$.

The prescaler value of 8 is selected with KA low, KB low and KC high as per Table 1. We then program the top digit of the counter using J4, which we set high, to 1. The remaining three digits are set to 2, 5 & 0, as shown in Table 2.

One 'gotcha' when setting up this counter is that, while the thousands digit for the counter is set using low-numbered inputs (J2-J4), the hundreds digit is set using the highest-numbered inputs (J13-J16). So the digits do not appear at the inputs in order, except in the mode when the prescaler can divide by up to 10.

Compact Frequency Divider



Figs.2 & 3: we recommend fitting all the SMDs on the top side first. Ensure all the ICs are orientated correctly and leave the SMA connectors, DC socket and headers until after you've populated the underside of the board. There are not as many components on the underside; just the comparator IC, passives and the dual diode.

slid them into place while heating that solder (to keep it molten). Removing the iron, we checked that all the leads were centred on their pads. If not, we reheated that solder joint and gently nudged the IC towards the correct position, rechecking each time.

Once the IC was positioned correctly, we soldered a couple more pins, then spread a thin layer of flux paste along both rows of pins, loaded the soldering iron tip with some solder and dragged it along the pins. Each one took up the right amount of solder, making quick work of all the joints.

Only a few joints got too much solder, resulting in a bridge to an adjacent pin. We removed the bridges using a bit more flux paste and an application of clean solder wick.

You could use a slightly different technique, where you clamp the device in the correct location using a clothes peg, haemostat clamp or similar, tack it down, then solder the remaining pins. That technique involves more set-up time but less trial-and-error.

Once the ICs are in place, you can solder the remaining three-lead and two-lead components with a similar technique. Just make sure you let one joint solidify (which can take a few

seconds) before making the other, or you could end up pushing the parts out of position.

With all the parts in place, clean the board with some flux cleaner (or pure alcohol if you don't have a specific flux cleaner), let it dry and inspect all the solder joints to ensure you haven't missed any imperfect/incomplete joints or bridges. Then flip the board over and solder the parts on the other side using a similar technique.

There is just one chip (IC6) on the underside, plus one dual diode in a three-pin SOT-23 package and 10 passives (resistors & capacitors). Take care with the orientation of IC6; its pin 1 goes towards the nearest PCB edge.

Some parts are close to IC6, so it's best to solder IC6 first, then the components right next to it, followed by those further away. Again, when finished, clean off the flux residue and inspect your work.

Finally, flip the board back over and solder the SMA connectors, the three-pin header for LK1, plus whichever of CON5 and CON6 you will be using. If leaving CON5 off, you will need to solder the short wire link shown in red in Fig.2 and the PCB silkscreen, or the board won't get power.

Note that you could leave SMA

connector CON2 off if you don't need or want the 1MHz output.

Testing

The board should draw under 20mA when powered up. If you have a current-limited bench supply, set it to 6V and at least 30mA and connect it to CON5 or CON6. If it goes into current limiting, switch it off and check for faults. If you don't have a bench supply, use a regular DC supply fed through a DMM set to measure milliamps and switch off if the current shoots up when you power it up.

Lacking such a supply, you just have to YOLO it: plug a suitable power supply in and check if LED1 lights. If it doesn't, unplug the cable and try to figure out why. If it does, proceed with the following checks.

Assuming the current draw is OK, check the voltage between the shell of one of the SMA connectors (ground) and the large tab of REG1. It should be close to 5V. If it is below 4.75V or above 5.25V, check the soldering on REG1 and its adjacent bypass/filter capacitors.

If it isn't drawing any current and the LED is off, that probably means that Q1 is not conducting. You can verify that by measuring the voltage between

Table 1 – 74HC4059 modes (• must be set up with Master Preset mode first)

KA	KB	KC	Prescaler ratio	Preset inputs	Counter thousands digit	Preset inputs	Maximum count
1	1	1	2:1 to 1:1	J1	0-7	J2-J4	15,999 (17,331 extended)
0	1	1	4:1 to 1:1	J1, J2	0-3	J3, J4	15,999 (18,663 extended)
1	0	1	5:1 to 1:1 •	J1-J3	0-1	J4	9,999 (13,329 extended)
0	0	1	8:1 to 1:1	J1-J3	0-1	J4	15,999 (21,327 extended)
1	1	0	10:1 to 1:1	J1-J4	0	-	9,999 (16,659 extended)

Table 2 – our 74HC4059 configuration

KA	KB	KC	Prescaler preset (J1-J3)	Thousands (J4)	Hundreds (J13-J15)	Tens (J9-J12)	Units (J5-J8)
0	0	1	000 (0)	1 (1)	0010 (2)	0101 (5)	0000 (0)

Constructional Project

your supply negative and the shells of the SMA connectors. There should be very little difference. If you measure the full supply voltage, check that you've applied power with the correct polarity. If you have, there is a fault around Q1/ZD1.

Finally, assuming the current draw is OK and the 5V rail is close to 5V, feed a signal with a known frequency into CON1 and check for $\frac{1}{10}$ th that frequency at CON2 (if you didn't fit CON2, you can probe its centre pin). If that checks out, apply 10MHz to CON1 and look for a 1Hz output at CON3. If it's missing, make sure JP1 is inserted in one of the two possible positions.

Remember that, depending on your test instrument, it could take several seconds to register a reading of such a low frequency.

If the board isn't behaving, common problems to look for are solder bridges, pins where the solder hasn't adhered to the PCB pad below, or incorrectly orientated ICs (we did warn you!).

Usage

There isn't much to it: connect your reference signal source to CON1 and feed the output at CON3 to your GPS clock(s) or other devices needing 1Hz pulses. Move JP1 if necessary to get the desired duty cycle, although almost any device expecting a 1PPS signal should work in either position.

We suggest housing the board in a small diecast aluminium box with the case connected to circuit ground to minimise EMI pickup. However, we tested it as a 'bare board' and it performed well in our lab. The SMA connectors are arranged along one edge, so you can mount the board such that they project through holes in the case, then add a chassis-mounting DC socket wired to CON6.

The four corner mounting holes will provide a convenient way to attach the board to the inside of such a box. If you need to make the board as small as possible, the tabs those holes are on can be cut off with a hacksaw or similar (but don't breathe the resulting dust

and cut them outdoors or in a well-ventilated area).

Most oscilloscopes, spectrum analysers, high-end frequency counters etc will have a pretty accurate 10MHz output; it's usually specified as something like ± 1 ppm. That isn't as good as a GPS-disciplined oscillator but it's still very precise. You will likely need a BNC-to-SMA cable to make this connection. You may need a second similar cable for the 1Hz output, depending on where it's going.

Lacking that, some newer DSOs have a waveform generator output that can generate a 10MHz sinewave or square wave (either is suitable). They tend to have quite a bit less stability and more jitter than a 10MHz reference output. However, an actual GPS 1PPS signal has jitter, so if you are using this board to emulate such a signal, you generally needn't worry too much about it.

You can also get connectors that break a BNC connection out to screw terminals if you're going to feed the 1PPS signal to pin headers or similar.

If you want to feed the 1Hz output of this board to multiple clocks or other devices, given its low frequency and the fact that most 1PPS inputs will have a high impedance, you will probably just need to 'fan it out'. You could even omit CON3 and solder wires directly to its pads. We mainly provided the SMA connector for convenience in hooking it up to prebuilt test equipment.

If you need to split the 10MHz output of your test equipment to go to multiple locations, consider building our Frequency Reference Signal Distributor (April 2021). However, note that the design won't work on the 1Hz output without modification as it is AC-coupled at the input and outputs.

PE

Parts List – 10MHz Frequency Divider

- 1 double-sided PCB coded 04112231, 64 × 37.5mm
- 3 right-angle or vertical through-hole SMA connectors (CON1-CON3)
- 1 SMD USB Type-C power-only socket with six pins (CON4) •
- 1 PCB-mount DC barrel socket (CON5) •
- 1 2-way polarised header, 2.54mm pitch (CON6) •
- 1 3-pin header, 2.54mm pitch (JP1)
- 1 jumper shunt (JP1)

• omit any of these power input connectors that are not needed

Semiconductors

- 3 (CD)74HC4017(M96) high-speed CMOS Johnson decade counters, narrow body SOIC-16 (IC1, IC2, IC4)
- 1 (CD)74HC4059 high-speed CMOS programmable divide-by-N counter, wide body SOIC-24 (IC3)
- 1 MC74VHCT50A hex CMOS non-inverting buffer, SOIC-14 (IC5)
- 1 TLV3501AID rail-to-rail high-speed comparator, SOIC-8 (IC6)
- 1 AMS1117-5.0 or compatible 5V 1A low-dropout regulator, SOT-223 (REG1)
- 1 AO3400 30V 5.8A N-channel logic-level Mosfet or equivalent, SOT-23 (Q1)
- 1 SMD LED, SMA/M3216/1206 size, any colour (LED1)
- 1 BZX84C5V6 5.6V 1% tolerance zener diode, SOT-23 (ZD1)
- 1 BAT54S dual series schottky diode, SOT-23 (D1)

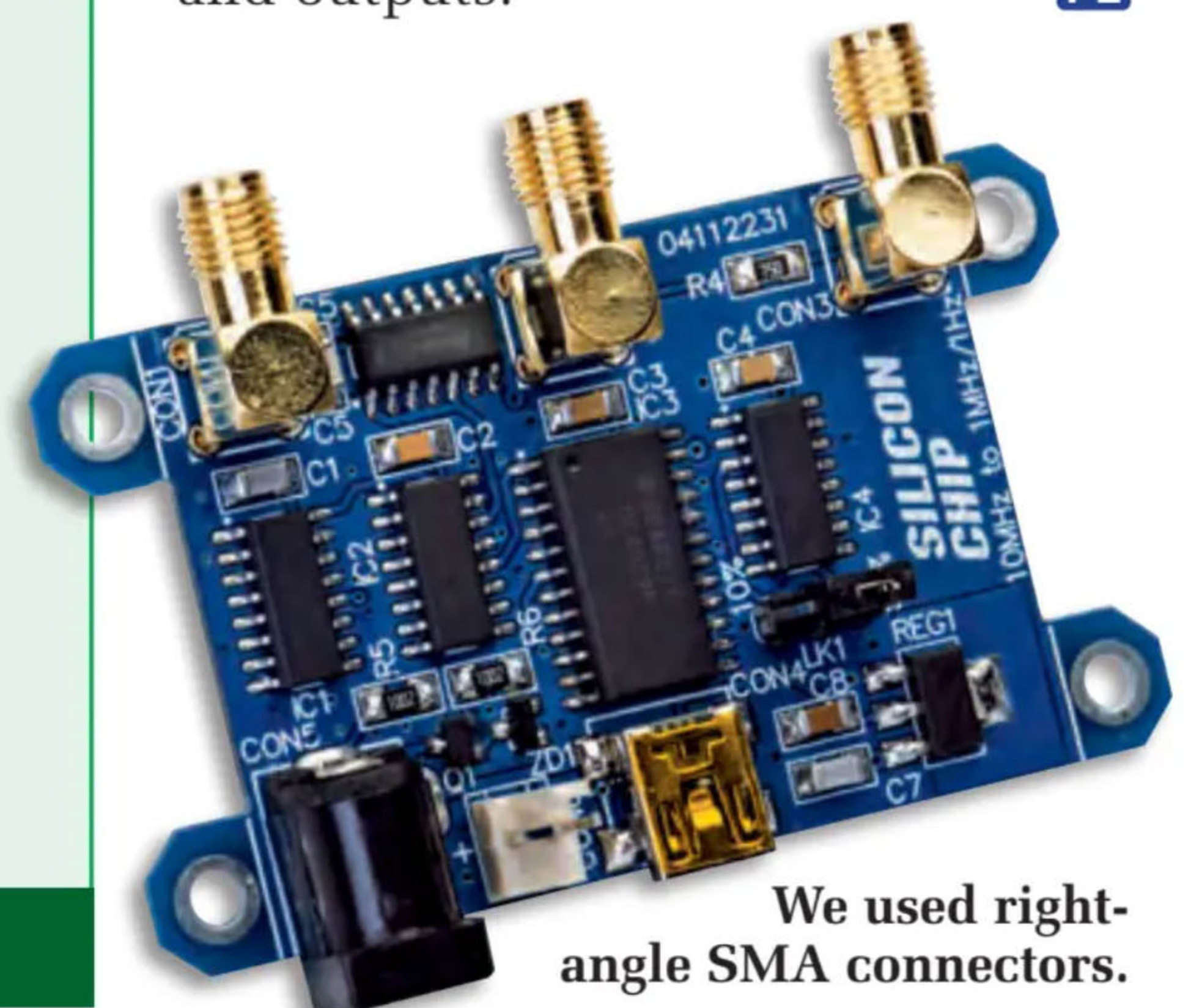
Capacitors (all SMD M3216/1206 size 50V X7R)

- 1 1 μ F 8 100nF 1 1nF

Resistors (all SMD M3216/1206 size 1%)

- 1 10M Ω
- 1 47k Ω
- 4 10k Ω
- 2 5.1k Ω (only needed if USB socket is fitted)
- 1 1k Ω
- 1 220 Ω
- 3 49.9 Ω , 51 Ω or 75 Ω (to suit desired input/output impedance)

10MHz Frequency Divider kit from Silicon Chip (Cat SC6881): ~\$48 (£24) + P&P



We used right-angle SMA connectors.

Only a few years ago, most of us would have thought the technology behind life-sized AI-enabled human holograms was far away, in the distant future. Well, tomorrow must have come early, since I just saw people holding a conversation with such a beast!

I feel like I'm straddling multiple technological epochs and I'm not as limber as I used to be. For example, I was just reading a very interesting 6-part series on the Rise and Fall of Heathkit (<https://pemag.au/link/ac3c>) written by my friend Steve Leibson.

The 1960s and 1970s were golden years for Heathkits. These ranged from entry-level projects like building a crystal radio or an oscillator to practice Morse code, all the way up to advanced kits like ham radio transmitter/receivers and colour televisions. As Steve says, "Many engineers started their budding careers by building one or more kits made by the Heath Company".

I remember those far-off times as if they were yesterday. Meanwhile, I currently maintain a tenuous toehold in the present, in which artificial intelligence (AI) and machine learning (ML) applications are rampaging across the technological landscape. Things are currently moving incredibly fast in AI/ML space, where no-one can hear you scream. For example...

Deploying AI at the edge

It's common to hear the term 'edge' used in the context of embedded systems and the internet of things (IoT). This means different things to different people, but I'm talking about what I think of as the 'extreme edge', that is, the point where the internet 'rubber' meets the real-world 'road'.

We are currently sitting on the horns of a dilemma, and it's jolly uncomfortable, let me tell you. The crux of the problem is that almost everyone wants to add AI/ML to their embedded/edge applications and devices, but very few people have the necessary expertise to add AI/ML to their embedded/edge applications and devices.

Just saying "crux" causes (what I laughingly refer to as) my mind to reflect on Frank Zappa's playful and enigmatic line: "The crux of the biscuit is the apostrophe", from his 1974 album *Apostrophe (')*. This captures Zappa's genius for making

the mundane seem profound and the profound seem mundane. But we digress...

Recently, I had quite the fascinating conversation with the folks at DeGirum (degirum.ai). These chaps and chapesses have a cunning solution to this problem.

First, they have a "Model Zoo" that contains 1,000+ pre-trained AI/ML models ranging from things like people detection, face detection, age estimation, gender classification, and emotion classification (happy, sad, angry...) to vehicle detection, number (license) plate detection, pothole detection, fire and smoke detection, intruder detection and weapon detection.

These models, which can be used in isolation or in combination, address the needs of a vast range of markets and applications. That includes 'smart' cities, offices and homes; construction, infrastructure and industrial automation; retail analytics and digital signage... the sky's the limit.

The folks at DeGirum also have a hardware-agnostic PySDK (Python software development kit). The clever bit is that they have a collection of supported hardware platforms 'in the cloud'. Developers can combine their application software with DeGirum's pre-trained models and run everything remotely on different platforms to determine the optimal cost/performance tradeoffs.

Best of all, this is free! People pay based only on the number of PySDK runtime instantiations that are eventually deployed in real products in the field.

Closer than we think

Did you ever see the 2002 movie interpretation of H. G. Wells' *The Time Machine*? There's a scene set in a futuristic library where our hero converses with an AI-enabled hologram. You can find this moment on YouTube (<https://youtu.be/CQbkhYg2DzM>).

When I first saw this movie (the day it came out), this level of technology seemed to be something that was a long, long way in the future.

Well, I was just chatting with Edward Ginis, who is the chief technology officer (CTO) at a company called Proto (protohologram.com). They essentially have this technology in the here and now!

Here's how it works in a crunchy nutshell. First, they video someone talking for a few minutes. Let's say that someone is me. Based on this, their AI can extract my 'vocal signature' for use in a text-to-speech role.

It can also determine how my lips and the muscles on my face interact with the various spoken sounds. This includes the way I smile, frown, blink etc. It's also watching the way I move my hands and arms; my entire non-vocal communication or body language.

This is where things get extremely clever. Suppose you were lucky enough to be having a live conversation with me. While you were talking, although it wasn't my turn to speak, I would still be engaged in the conversation with my body language: smiling, frowning, nodding in agreement... that sort of thing. Also, my mind would be constantly evaluating my potential responses based on what you were saying.

Now suppose that you were talking to Proto's AI hologram version of me. While you were talking, the hologram would be performing its non-verbal communication activities. Also, it would be deciding how to respond.

This all happens in real-time (not like today's Alexa-like assistants that wait for you to finish talking, then upload what you've said, and then reply after a few seconds' delay). This means the AI hologram will respond immediately—sometimes before you've even finished talking—just like in a real conversation.

Also, all visual aspects of the AI hologram—lips, facial muscles, hand and body gestures—are synchronised with whatever it's saying.

And just what will it be saying? I'm glad you asked. I'll tell you about that in my next column. **PE**

Circuit Surgery

Regular clinic by Ian Bell

Topics in digital signal processing – windowed sinc filters

We are looking at various topics related to digital signal processing (DSP).

DSP covers a wide range of electronics applications where signals are manipulated, analysed, generated, stored or displayed as digital data but originate from and/or are converted to real-world signals for interaction with humans or other parts of the physical world.

Fig.1 shows the key elements of a generic DSP system with a signal path from an analog input via digital processing to an analog output. This does not necessarily represent every DSP system (not all have all the parts shown), but it serves as a reference for the various subsystems we will consider.

Last month, we started looking at the windowed sinc filter, which is a widely used type of digital filter. We discussed the “sinc” aspect of the filter’s name, which refers to the fact that its impulse response takes the form of the mathematical sinc function.

The impulse response is the output obtained from a system when the input is an impulse. For a sampled data system, this is a single input sample of value 1 at time zero, with an input of zero at all other times.

For a Finite Impulse Response (FIR) filter (Fig.2), the coefficients (the a_i values in Fig.2) are equal to the impulse response of the filter. So, if we know the impulse response of the desired filter, we have the coefficient values required to implement it.

As discussed last month, we can obtain the impulse response of a filter by performing an inverse Fourier Transform on the required filter frequency response. The result for an ideal low-pass filter (a ‘brick wall’ filter) is that the impulse response is a sinc function, hence the

importance of the sinc function in filter implementation.

We showed how to use a spreadsheet to calculate an example set of low-pass filter coefficients, normalising the DC gain to one.

Filter & simulation improvements

The sinc function has non-zero values over an infinite range, so a perfect filter would require an infinite number of coefficients. That is obviously not practical, so last month we simply used the relevant sinc function values up to a specific number of coefficients.

Truncating the coefficients like this does not necessarily produce the best results, which is where the “windowing” part of the filter name comes in. Before getting into that, though, we will continue our discussion on using LTspice to simulate these filters.

In previous articles, we used LTspice to implement digital filter maths for obtaining frequency response plots or transient simulations of just the filter, or the filter together with other subsystems such as sampling and reconstruction filters (see Fig.1). Our approach uses LTspice behavioural voltage sources to implement the delay, multiply and add operations shown in the filter structure of Fig.2.

Last month, we showed an FIR filter example with 25 coefficients, which required drawing 24 behavioural sources on the schematic to implement the delays. As a reminder, part of the schematic from last month is shown in Fig.3. This includes just five of the 24 delay sources, and the input signal source is also not shown.

We may need to investigate filters with many more coefficients, or circuits with multiple filters, but in doing so, it becomes increasingly cumbersome to draw

the schematic. We need a better way to create large filters.

SPICE netlists

Although we usually draw schematics when using LTspice, the simulator actually uses a text description of the circuit, called a netlist, which is generated from the schematic drawing. SPICE simulators go back to the 1970s, before the days of high-resolution graphical user interfaces, so text has always been the format for circuit input.

The libraries of component models provided with LTspice or via downloads are also primarily text-based netlists or model statements.

You can view the netlist of a schematic from the main menu using View → SPICE Netlist. For the circuit from last month, the netlist will be similar to that shown in Fig.4, although the lines may be in a different order.

Fig.4 only shows the first and last three behavioural sources used to implement the delays. Still, it should be clear that the regular structure of the circuit leads to a repetitive form for the code relating to these sources in the netlist.

This means it is reasonably easy to create the netlist of a larger filter without

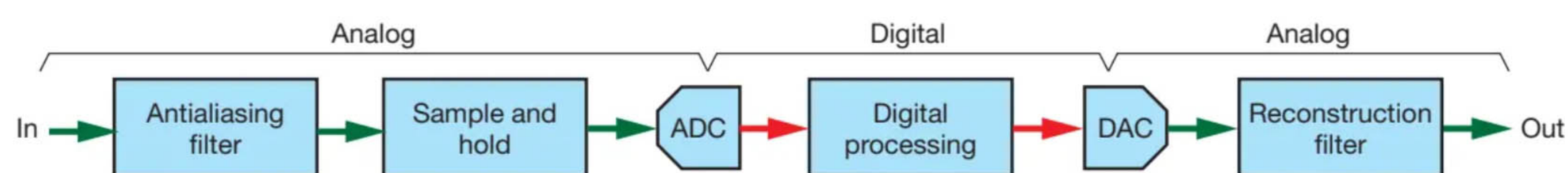


Fig.1: a generic digital signal processing (DSP) system structure.

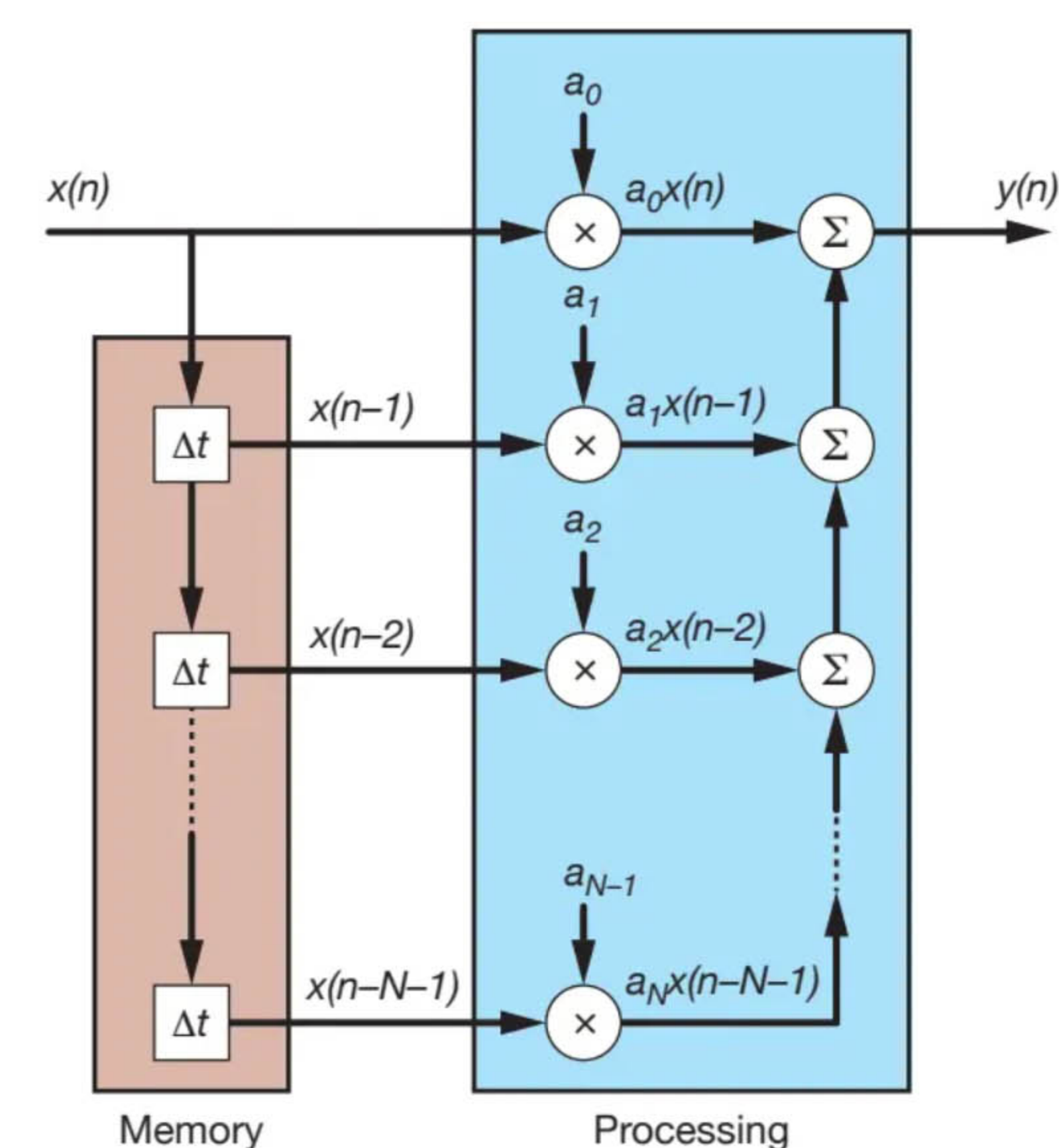


Fig.2: the structure of a finite impulse response (FIR) filter.

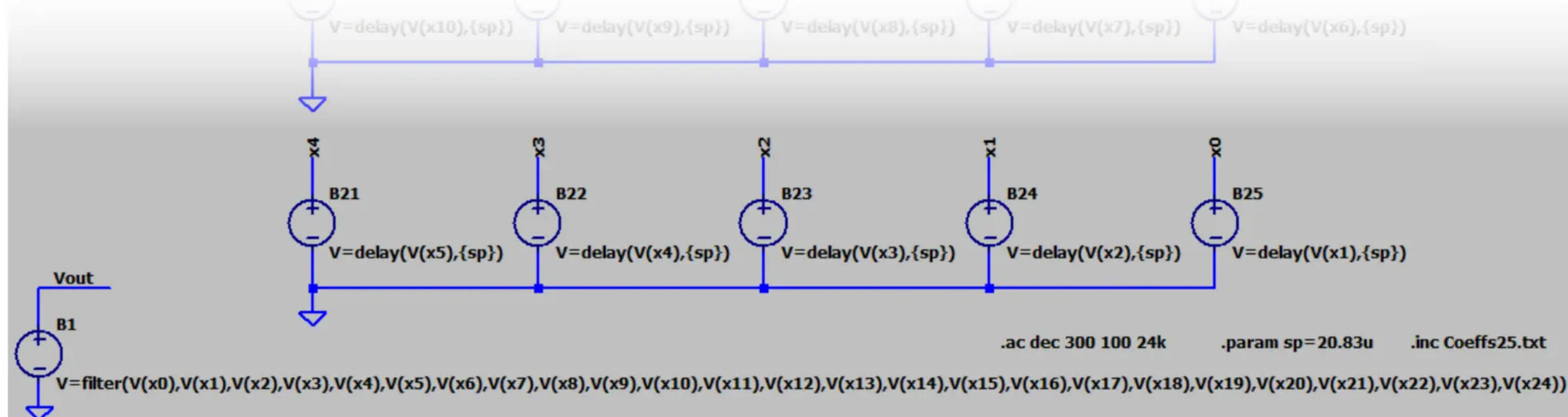


Fig.3: part of the filter schematic from last month showing 5 of the 24 delay sources. The input signal source is also not shown.

drawing a schematic. A single line can be pasted many times and the index numbers of the signals edited on each line, or software code, or a spreadsheet can be used to generate the text.

The filter output function and coefficient table are also required for a complete implementation (remember, some of the filter definitions are in the “include file” [Coeffs25.txt] in the example Figs.3 & 4). However, these also have a well-defined structure that is straightforward to generate using code or a spreadsheet.

Subcircuits

The best way to use a netlist for the filter implementation is to create a subcircuit containing the netlist. A SPICE subcircuit is a netlist that can be used like a component in other circuits. It can be linked to a symbol and included in other schematics, or simply referenced via text statements.

Libraries of symbols linked to subcircuit-based models of real devices, such as op amps, are included in the LTspice download.

Defining a subcircuit involves two SPICE directives (commands): **.subckt** and **.ends**. The general format is:

```
.subckt name p1 p2 p3 ...
    circuit netlist
.ends
```

Here, **name** is the name of the subcircuit and **p1, p2, p3** etc are the pin names (the input/output connections of the subcircuit). Between these two directives is a normal SPICE netlist defining the structure of the subcircuit.

To use a subcircuit, it can be referred to using a text statement in another netlist or on a schematic. Subcircuits are components when used in other circuits, so the syntax follows the same general structure as other components. Subcircuit instance (individual component) names start with an X, similar to how resistor names start with an R and capacitors with a C.

The syntax for using a subcircuit is:

```
Xxxxxx n1 n2 n3 ... name
+ param_name1=param_value1 ...
```

Here, **xxxxx** is the name of the individual

```
* D:\LTspice\DigitalFilter\DSP-SincFilters-fig9.asc
V1 x24 0 PULSE(0 1 2m 1u 1u 2m 4m) AC 1
B2 x23 0 V=delay(V(x24),{sp})
B3 x22 0 V=delay(V(x23),{sp})
B4 x21 0 V=delay(V(x22),{sp})
:
B23 x2 0 V=delay(V(x3),{sp})
B24 x1 0 V=delay(V(x2),{sp})
B25 x0 0 V=delay(V(x1),{sp})
B1 Vout 0 V=filter(V(x0),V(x1),V(x2),V(x3),V(x4),...)
.param sp=20.83u
.ac dec 300 1 24k
.inc Coeffs25.txt
.backanno
.end
```

Fig.4: an abbreviated SPICE netlist of a schematic.

```
.subckt lp_5k_48k_rec_25 inp outp

.func a(i) {table(i,0,0.0260921095,1,0.02258210425,...)}

B1 x23 0 V=delay(V(inp),{sp})
B2 x22 0 V=delay(V(x23),{sp})
B3 x21 0 V=delay(V(x22),{sp})
:
B22 x2 0 V=delay(V(x3),{sp})
B23 x1 0 V=delay(V(x2),{sp})
B24 x0 0 V=delay(V(x1),{sp})

B25 outp 0 V=a(0)*V(inp)+a(1)*V(x23)+ ... +a(24)*V(x0)

.ends
```

Fig.5: the LTspice subcircuit code for a sinc filter.

instance of the subcircuit (eg, X1 or Xfilter); **n1, n2, n3** etc are the nodes (wires) in the circuit which connect to the pins of the subcircuit (listed in the same order as the subcircuit definition); and **name** is the name of the subcircuit.

If the subcircuit netlist contains references to parameters (ie, the parameter name inside braces {**param_name1**}), the value(s) to use for the parameter(s) are specified after the subcircuit name.

Fig.5 shows the subcircuit code for a 10kHz low-pass filter with 25 coefficients (similar to the filter partly shown in Fig.3). The subcircuit description includes the delay sources discussed above, together with the table of coefficients and the filter output function that were in the include file previous examples (this was discussed in January 2025 *Circuit Surgery*).

Thus, the subcircuit file contains the

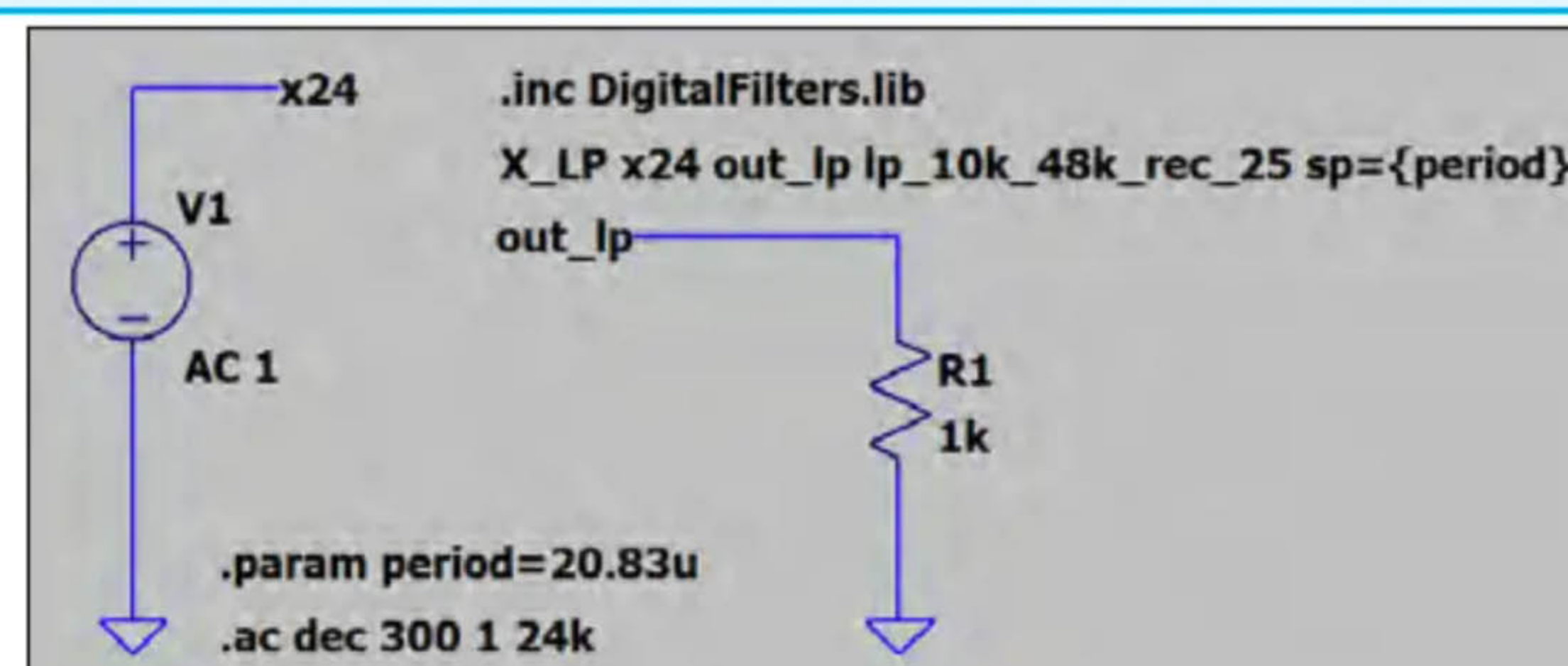


Fig.6: an LTspice schematic using a digital filter subcircuit.

complete filter definition. Only the beginning and end of the lines including the **.func** and **B25** statements in Fig.5 are shown here due to their lengths.

Simulating the subcircuit

A subcircuit can be referenced on a schematic as shown in Fig.6. The code in Fig.5 is in the text file **DigitalFilters.lib**, which is referred to as an include file on the schematic. The extension **.lib** (library) is commonly used for LTspice subcircuits, which are often found in component libraries, but it is just a text file and can use other extensions.

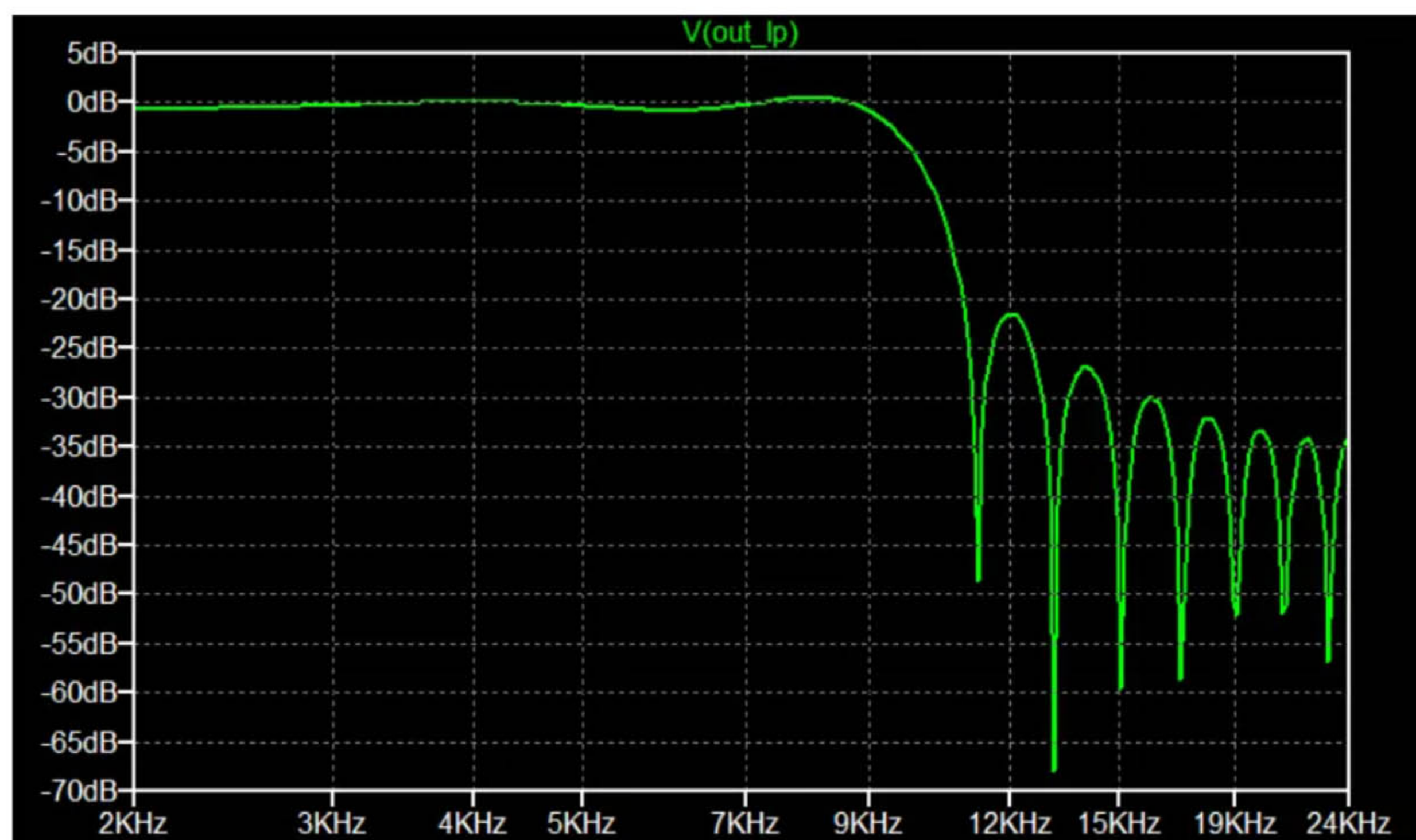


Fig.7: the simulation results from the circuit in Fig.6 (dB/log plot).

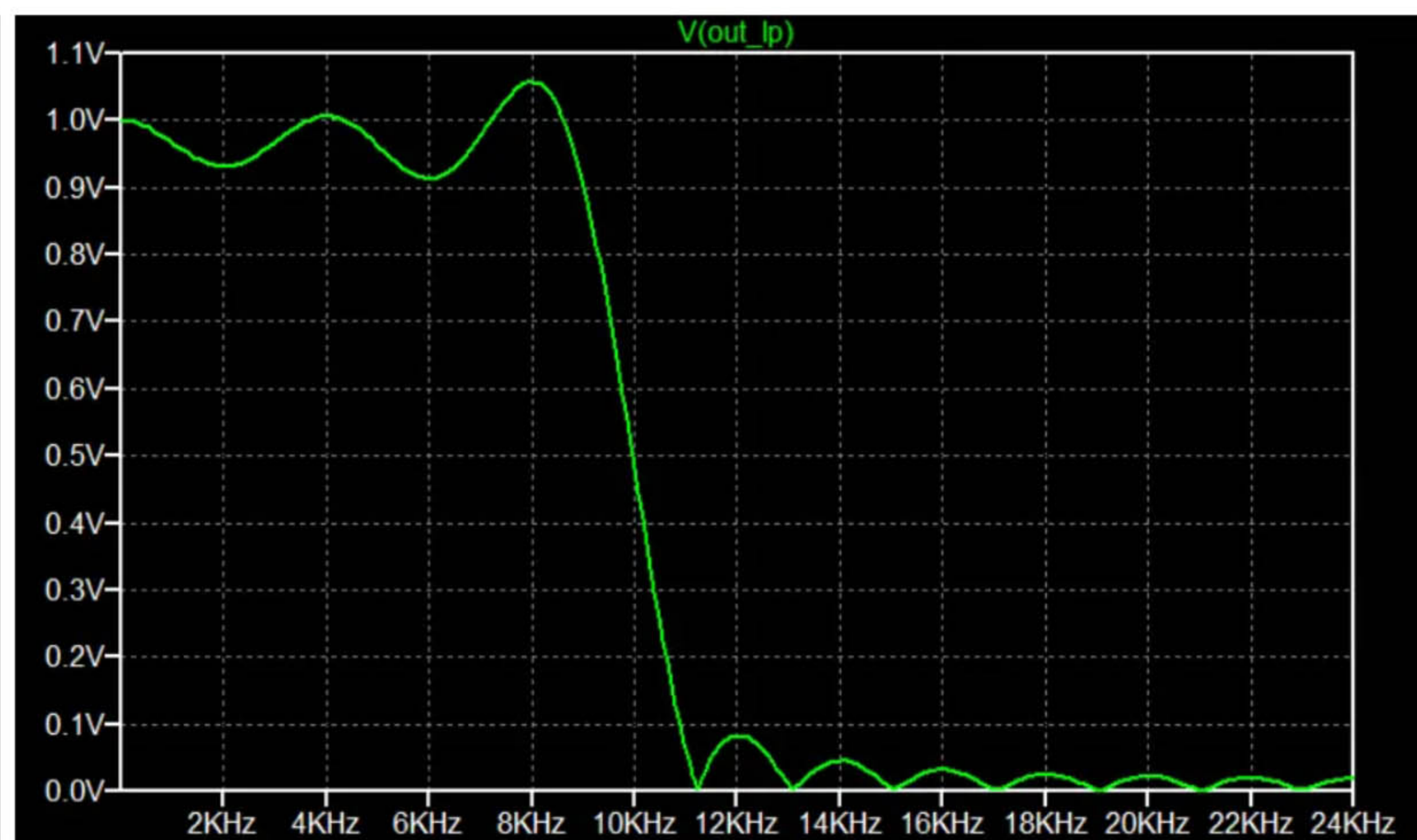


Fig.8: the simulation results from the circuit in Fig.6 (linear plot).

Subcircuit/library files may contain multiple subcircuit definitions. The function of the circuit in Fig.6 is essentially the same as the example from last month partly shown in Fig.3, although the cutoff frequency is different. Consider the following statement on the schematic:

```
X_LP x24 out_lp
+ lp_10k_48k_rec_25 sp={period}
```

This statement creates an instance of the **lp_10k_48k_rec_25** subcircuit defined in the **DigitalFilters.lib** file, with the component name **X_LP**. It is connected to nodes **x24** (filter input) and **out_lp** (output). The parameter **sp** (sample period) used inside the subcircuit is set equal to the value of the parameter **period** defined on the schematic using **sp={period}** in the subcircuit statement.

The **DigitalFilters.lib** file must be in the same folder as the schematic; otherwise, the full path would be required in the **.inc** statement.

The 1kΩ resistor (R1) on the filter output is not necessary for the simulation to run, but serves as a focus for the output signal, given that there is no circuit symbol for the filter itself. Of course, in a real circuit, the next stage would present some input impedance, which this resistor could represent.

As the output of the filter is an idealised voltage source, the resistor value is not particularly important in this example.

The results of the simulation of the circuit in Fig.6 are shown in Figs.7 & 8

(decibel/log frequency and linear plots, respectively). The cutoff frequency of 10kHz occurs at a gain of 0.5, or -6dB, not the more commonly used definition of -3dB.

In general, the definition of cutoff is somewhat arbitrary on a smooth response curve and may be different for different types of filter. The mathematics of coefficient calculation for sinc filters naturally maps the desired cutoff to the 0.5 gain point.

Also, the stopband and passband responses are symmetrical about the 0.5 point. If you imagine the peaks in the stopband at 12kHz and 16kHz in Fig.8 going negative, this may be easier to understand. So a gain of 0.5 is a natural choice for the cutoff frequency definition.

The schematic in Fig.6 does not have a circuit symbol for the filter. The include file and subcircuit component statements on the schematic are written directly into the netlist, so they work fine. However, if a symbol is preferred, it is possible to create one.

Windowing

The response of the sinc filters we have investigated so far is not particularly good. There is significant ripple (gain variation) in the passband (see Fig.8) and the stopband attenuation is poor, being only -21dB at its worst level (see Fig.7).

This situation is not improved by increasing the number of coefficients, although as discussed last month, this will improve the sharpness of cutoff (reduce

the transition bandwidth). The problem is caused by the fact that the sinc function of the impulse response extends to infinity, but we are only using part of this function.

For a filter with N coefficients, we are effectively truncating or windowing the sinc function to only those values in the $\pm(N-1)/2$ range. Mathematically, this is equivalent to multiplying by a rectangular function, which is 1 in the range $\pm(N-1)/2$ and zero elsewhere.

Fig.9 shows the impulse response for a 5kHz low-pass filter with 48kHz sampling rate. This was obtained using a spreadsheet similar to the one discussed last month. Fig.10 shows a rectangular window function for $N=25$ and Fig.11 shows the result of multiplying the impulse response by the rectangular window.

There is an abrupt cutoff at the end of the windowed impulse response, which tends to impact negatively on filter performance.

The rectangular widow occurs by default if we just truncate the sinc function to use a specific number of coefficients in the filter. However, we can use other functions that will reduce the sinc function coefficients towards the edges of the used range to make the transition at the ends less abrupt. Doing this can provide certain benefits.

One way to do this is to reduce the window value linearly from 1 at the centre to zero at the edges, as shown in Fig.12. This is known as a triangular or Bartlett

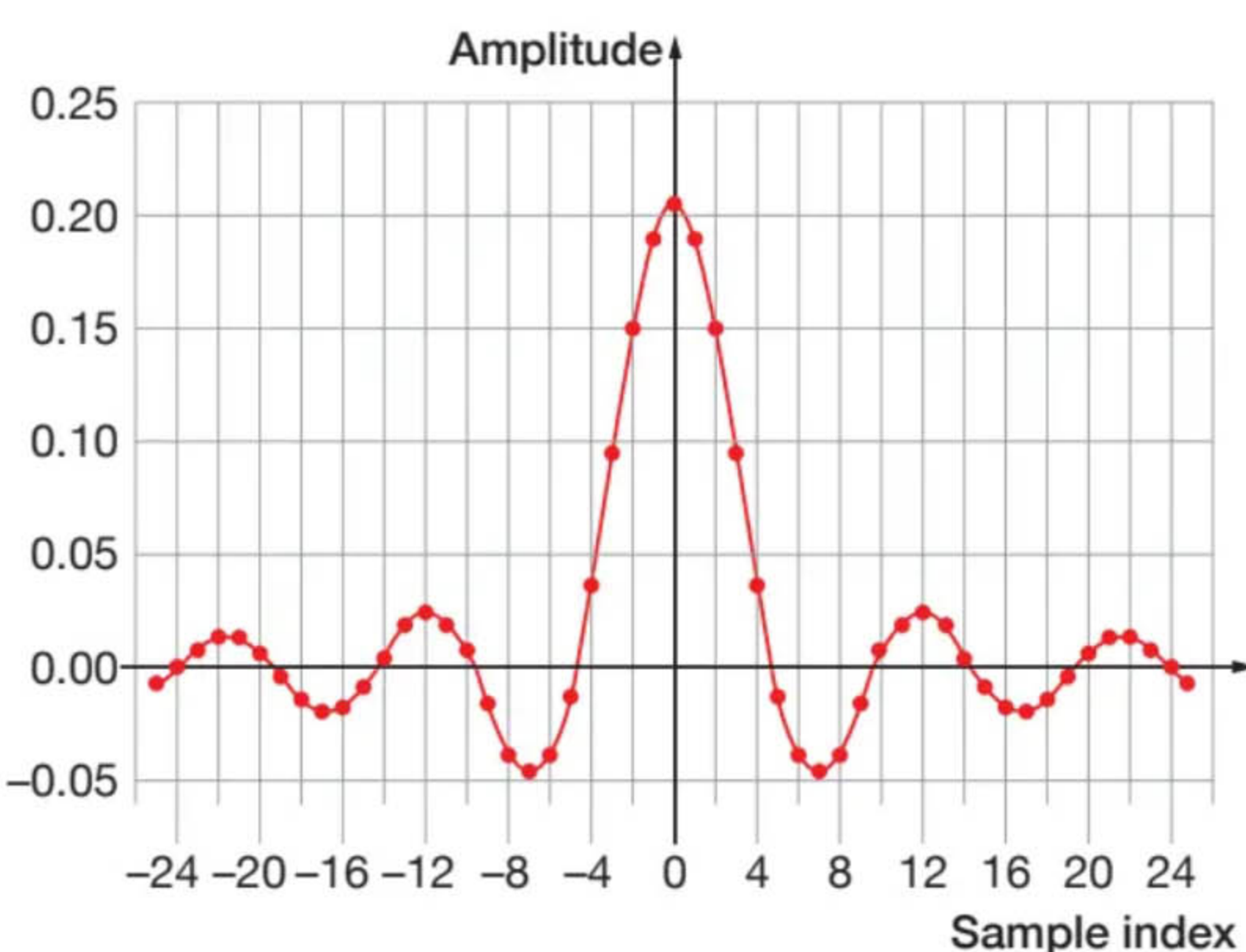


Fig.9: sinc impulse response coefficients.

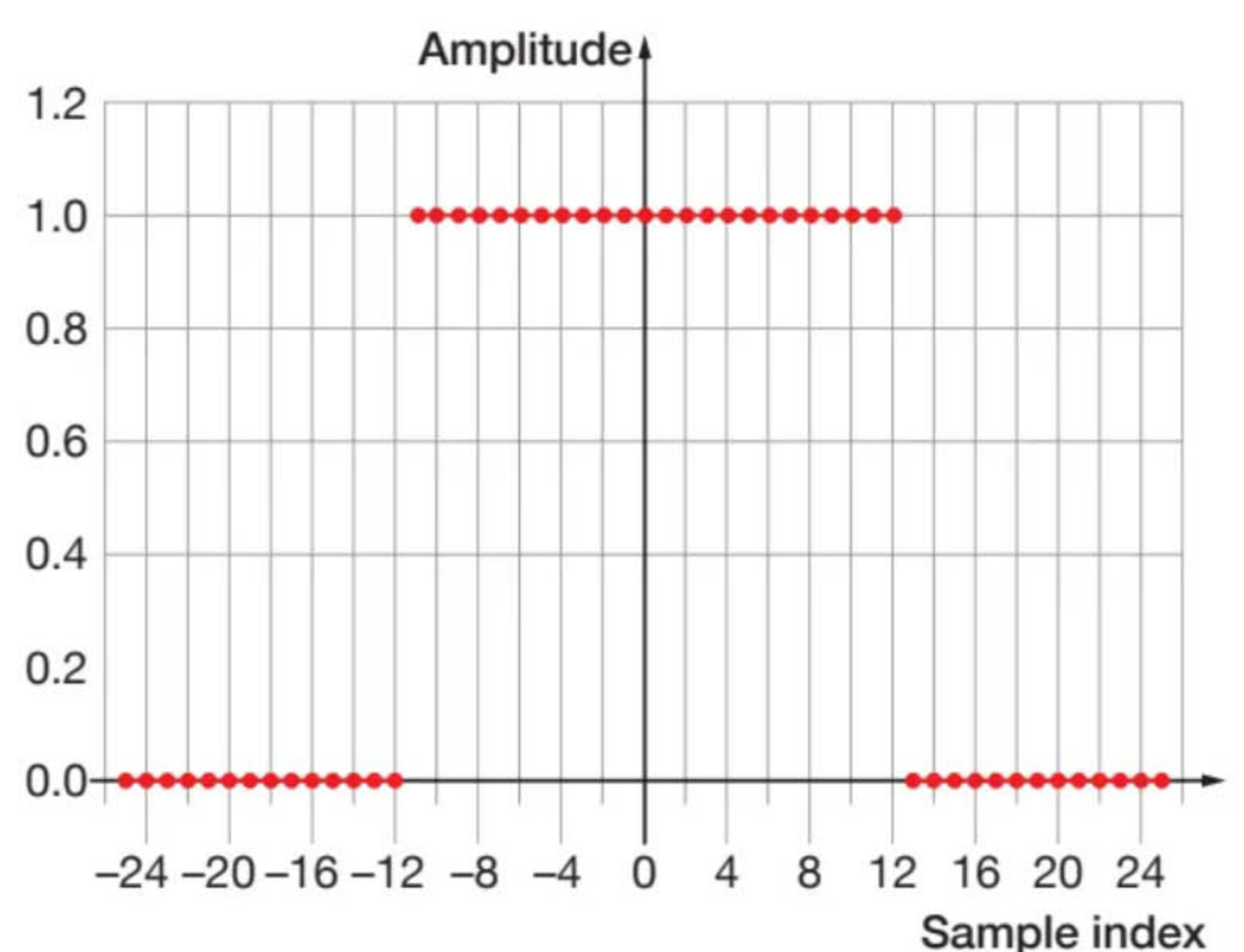


Fig.10: 25-coefficient rectangular window.

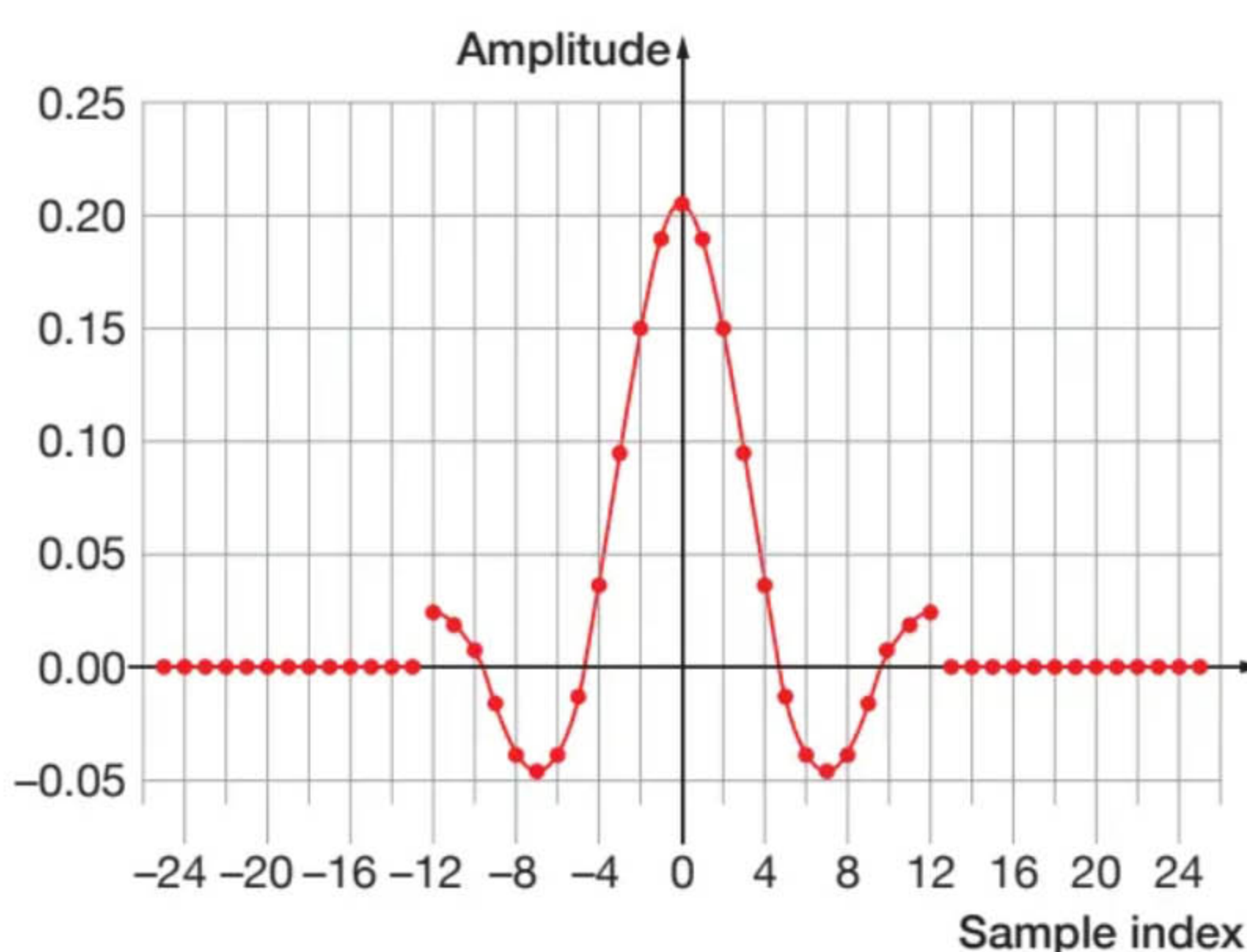


Fig.11: a plot of Fig.10 applied to Fig.9.

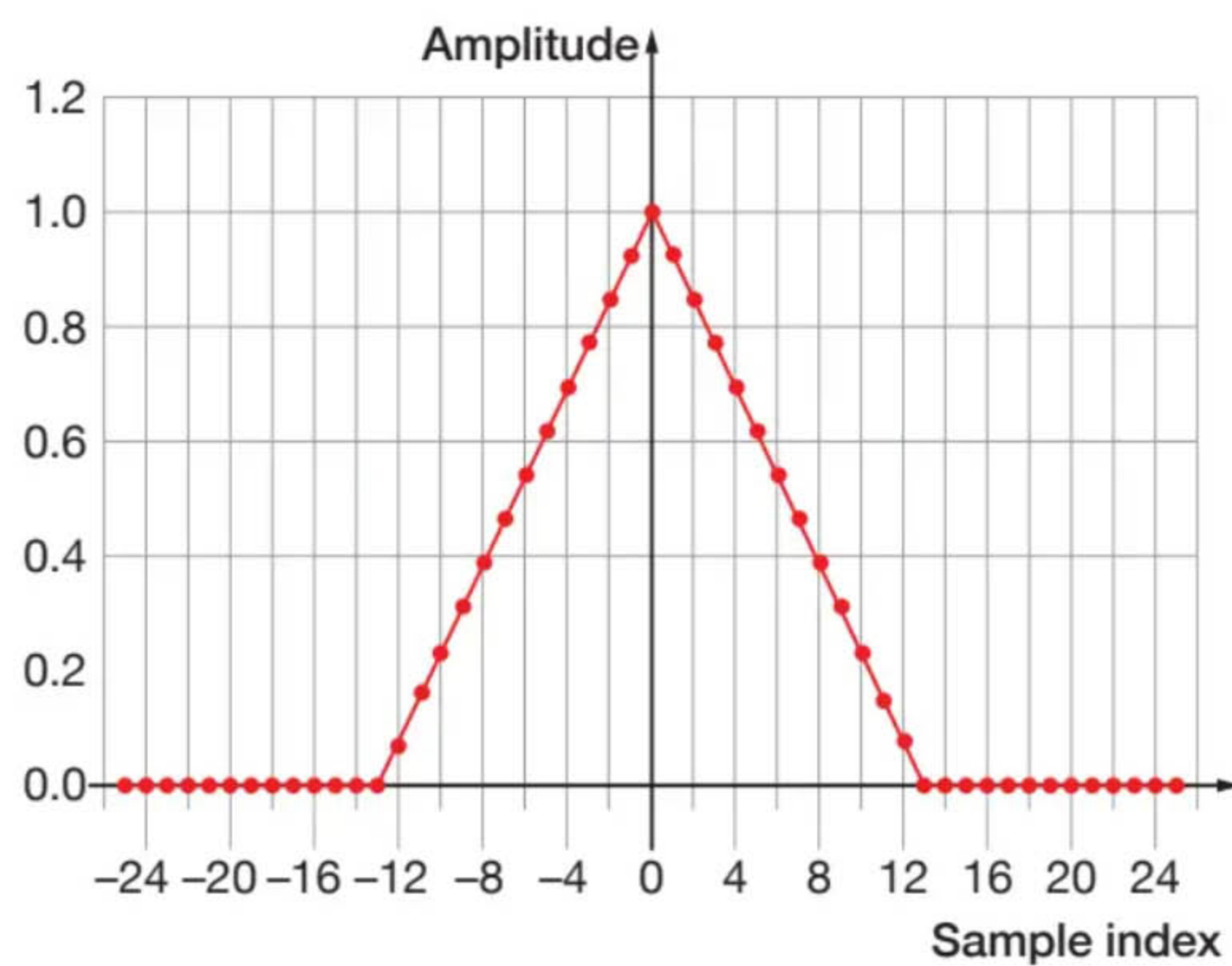


Fig.12: Bartlett window for 25 coefficients.

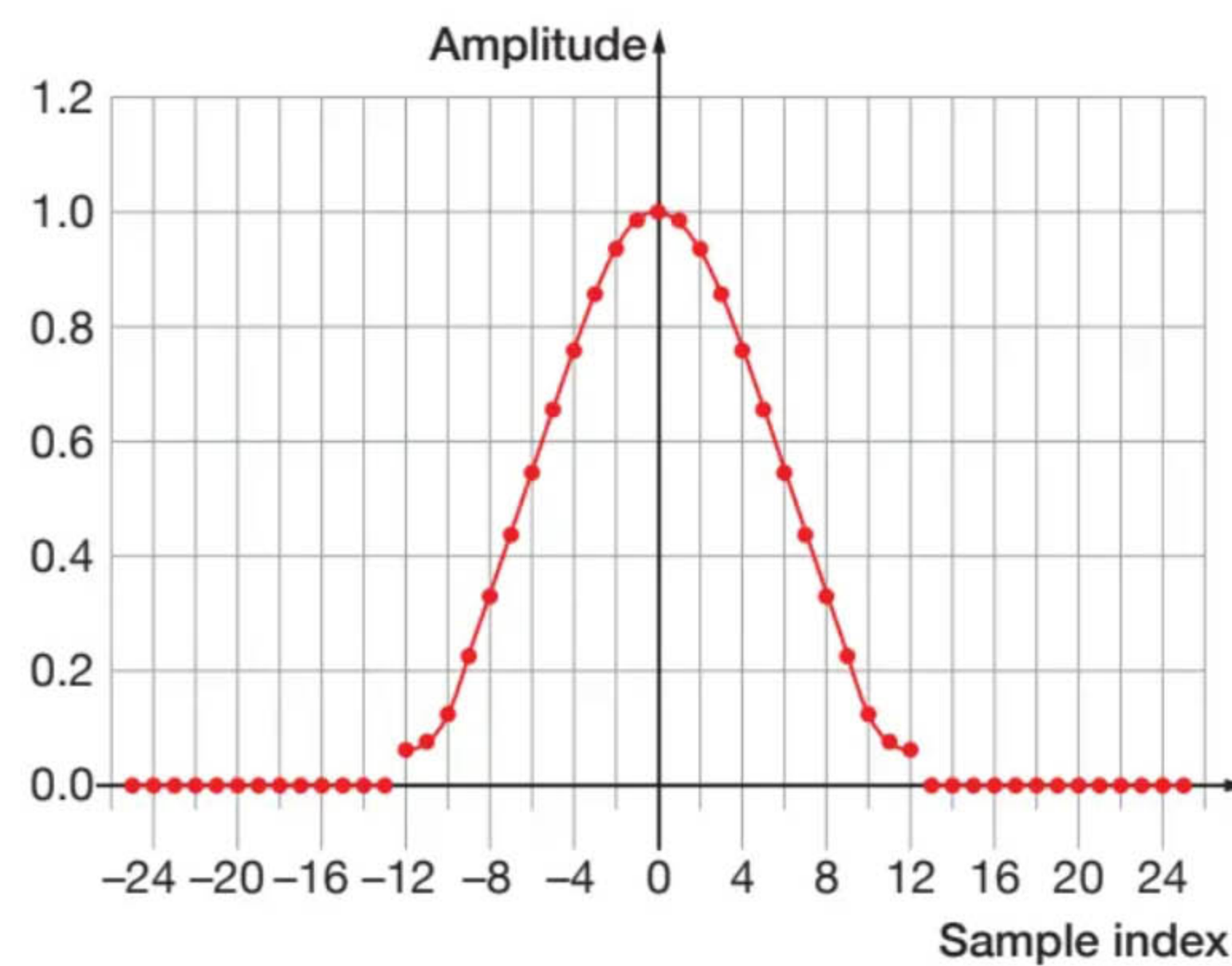


Fig.13: Hamming window for 25 coefficients.

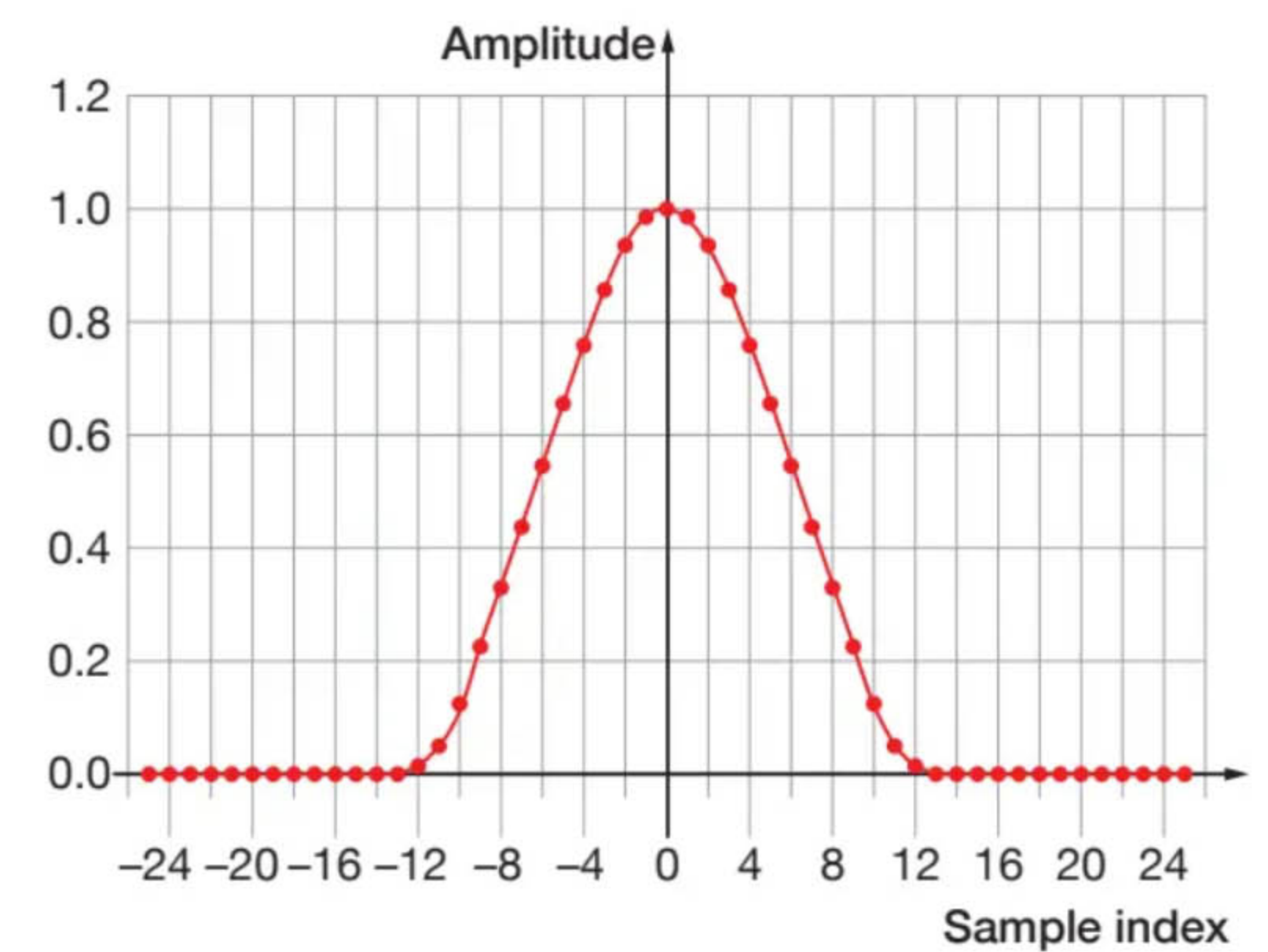


Fig.14: Blackman window for 25 coefficients.

window. It provides an improvement, but better performance can be obtained using more smoothly varying functions.

Window functions

There are a large number of possible window functions that can be used, but a few based on cosine functions are commonly used in FIR filter designs. These are the Hanning, Hamming and Blackman windows, the formulae for which are stated below.

These are written using the same symbols for sample index (i) and the total number of coefficients (N) as the formula for coefficients discussed last month. In all cases, the value is zero for index values outside the $\pm(N-1)/2$ range.

Hanning:

$$0.5 - 0.4 \cos\left(\frac{2\pi i}{N-1}\right)$$

Hamming:

$$0.54 - 0.46 \cos\left(\frac{2\pi i}{N-1}\right)$$

Blackman:

$$0.42 - 0.5 \cos\left(\frac{2\pi i}{N-1}\right) + 0.8 \cos\left(\frac{4\pi i}{N-1}\right)$$

These window functions are named after their originators, as is usually the case, although the Hanning window is based on work by Julius von Hann and was given the name “Hanning” by Ralph Blackman (of the Blackman window) and John Tukey. It is confusingly similar to the name of the Hamming window, which is attributed to Richard Hamming.

These three window functions have similar shapes. Figs.13 & 14 show the Hamming and Blackman windows, respectively. Fig.15 shows the Blackman window applied to the impulse response (filter coefficients) in Fig.9. Comparison with Fig.11 shows a significant reduction in abruptness of the transition at the ends of the index range (± 12).

Simulating windowed filters

Fig.16 shows an LTspice schematic for comparing the responses of the rectangular, Hamming and Blackman windows. The filters are all 5kHz low-pass types with a sampling rate of 48kHz. The approach is similar to that used in the circuit in Fig.6, and each filter is defined by a subcircuit like that shown in Fig.5.

All three subcircuits are included in

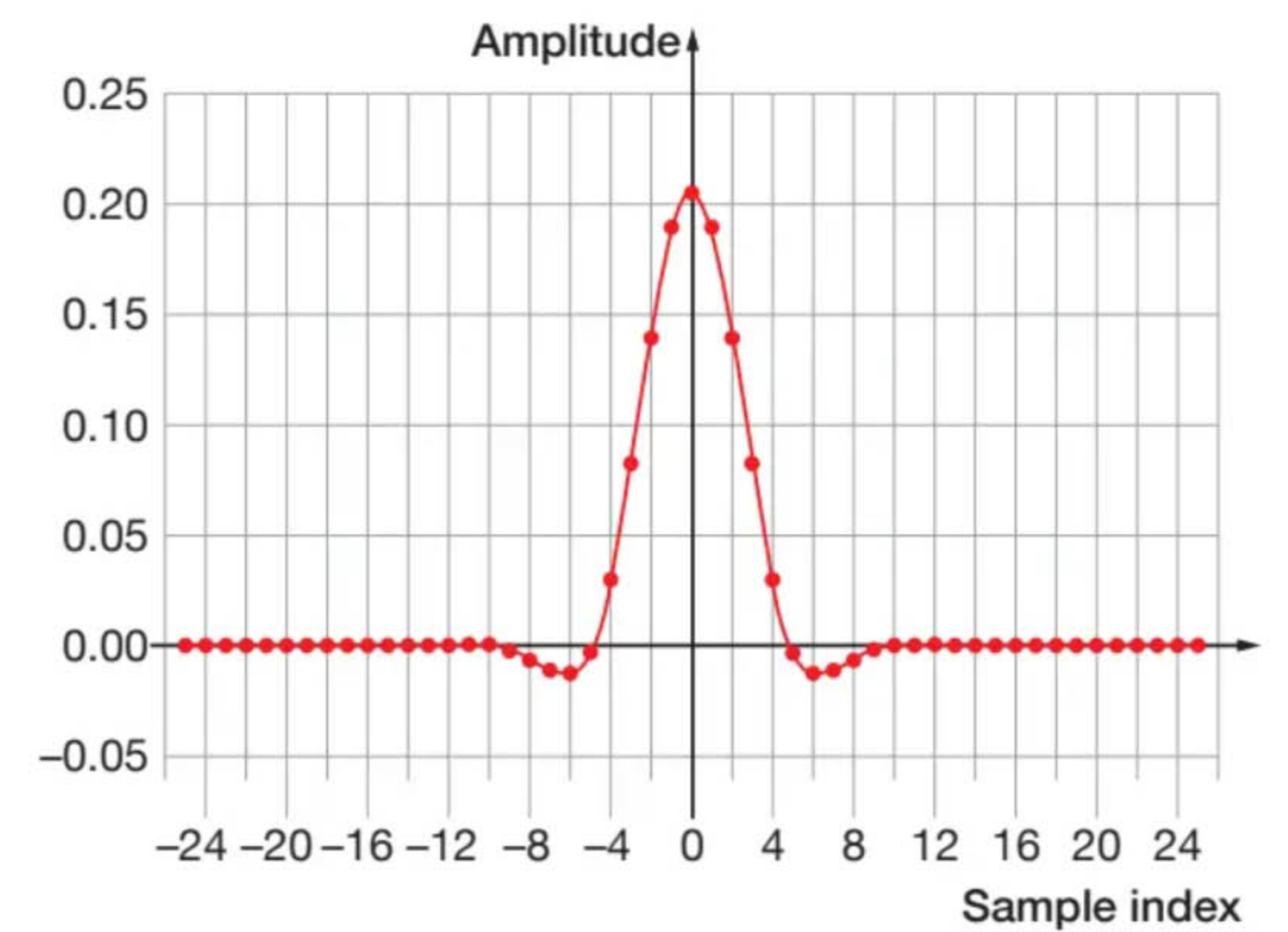


Fig.15: a sinc impulse response with the Blackman window in Fig.14 applied.

the file **WindowedFilters.lib**, and each one only differs from the code in Fig.5 by the coefficient values used in the table.

The coefficient values were calculated using a spreadsheet similar to the one discussed last month. The sinc function impulse response is calculated the same way as before and normalised to give the rectangular windowed filter coefficients. The window functions for the Hamming and Blackman windows are calculated using the formulae above over the same index range as the sinc function.

The raw sinc function is multiplied by the window function values to give the raw impulse response with these windowed filters. These are individually

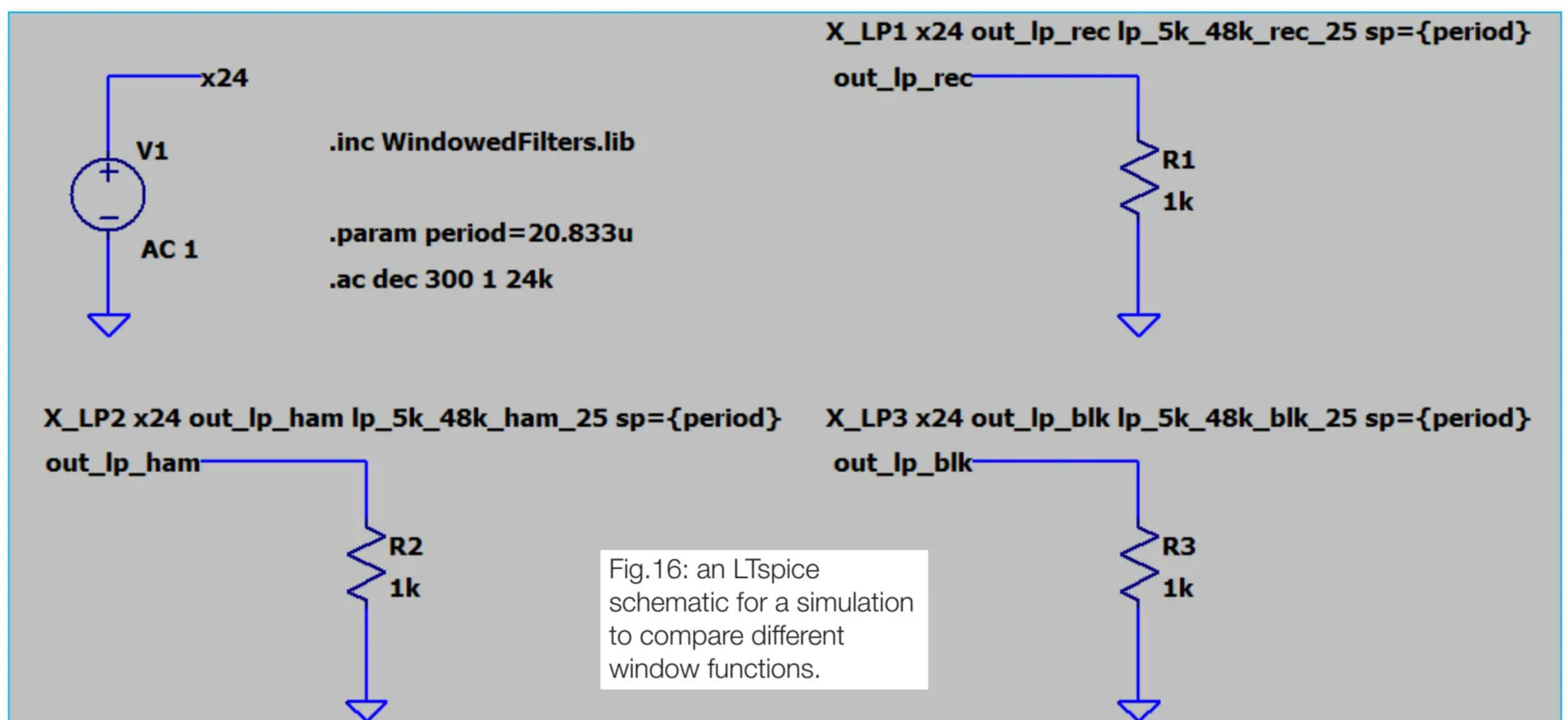


Fig.16: an LTspice schematic for a simulation to compare different window functions.

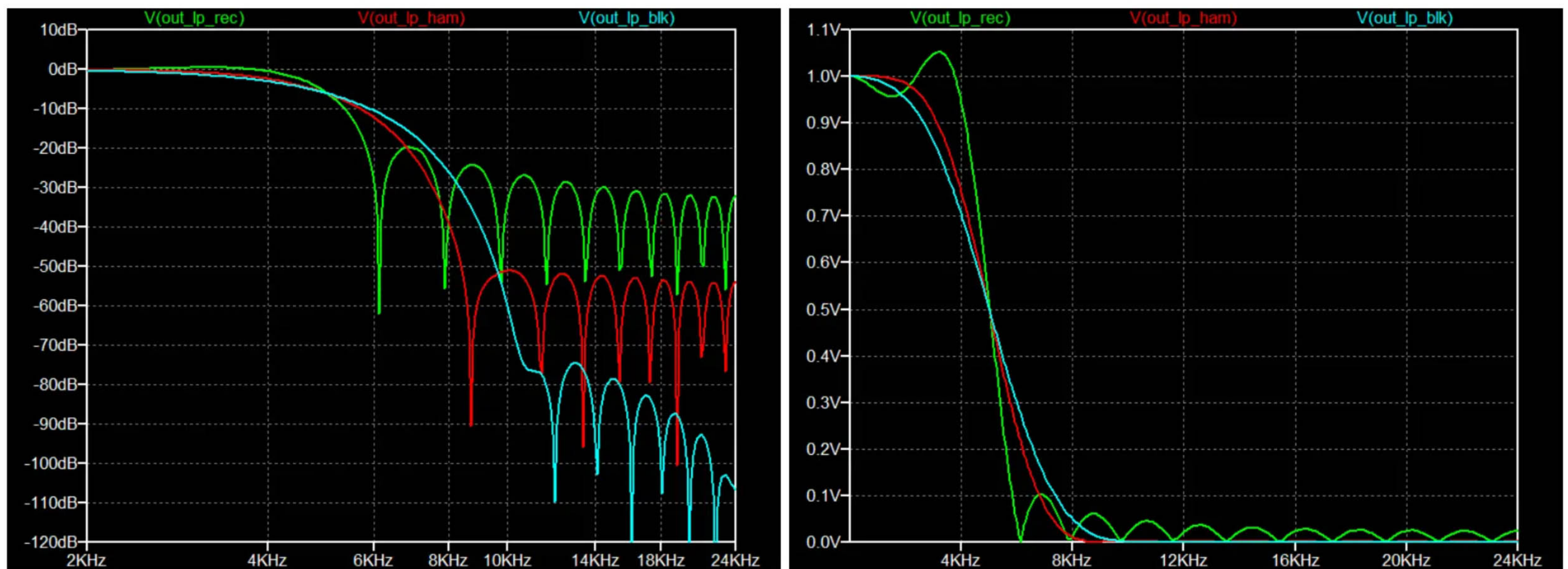
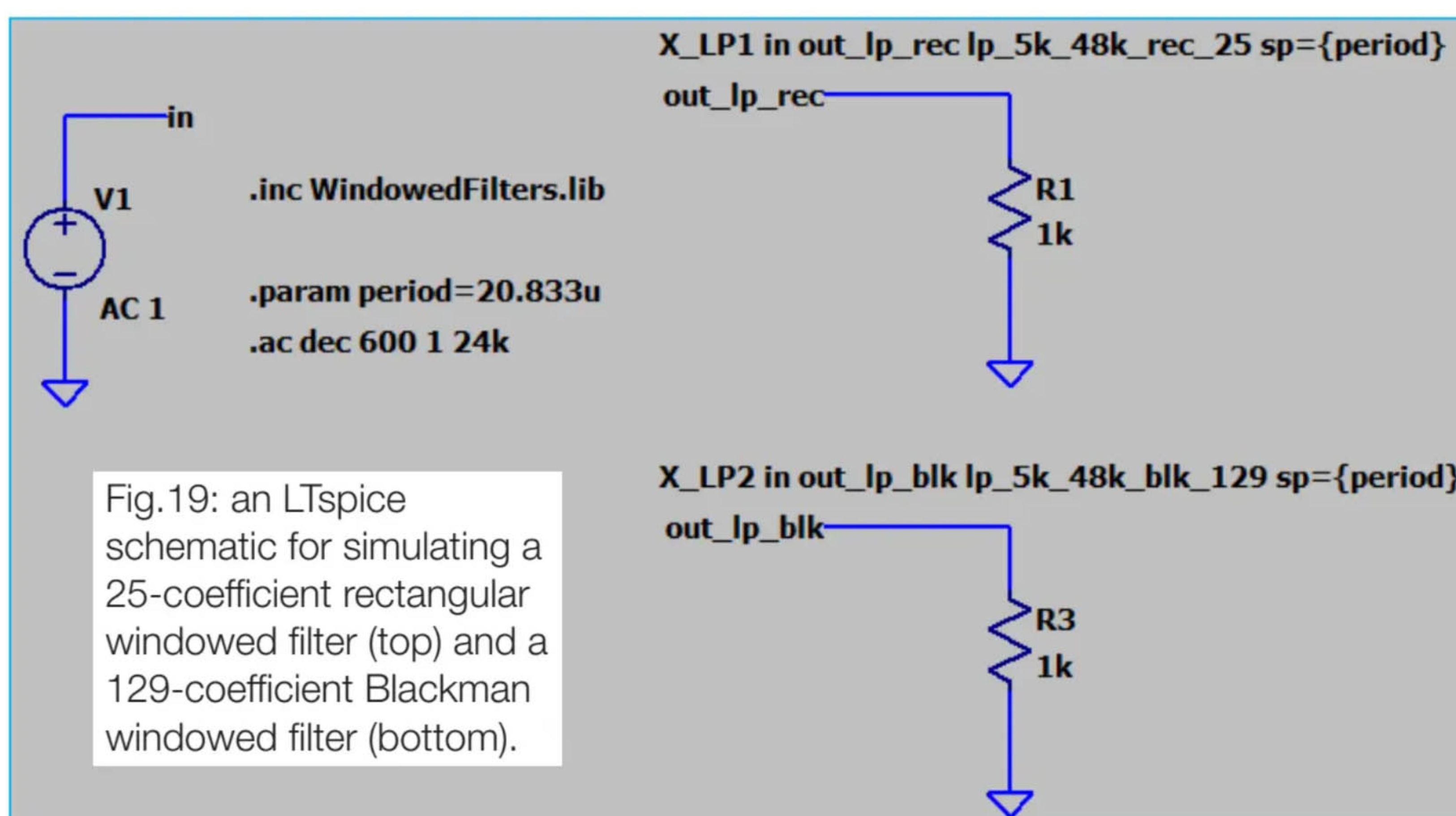


Fig.17: the simulation results from the circuit in Fig.16 (dB/log plot). Fig.18: the simulation results from the circuit in Fig.16 (linear plot).



normalised to obtain unity-gain versions of the windowed filters.

The results of simulating the circuit in Fig.16 are shown in Figs.17 & 18 (decibel/log frequency and linear plots, respectively). The green trace is the rectangular windowed filter, the red trace is the Hamming and the cyan one is the Blackman.

The Hamming and Blackman windows provide much better (worst-case) stopband attenuation at around 51dB and 74dB, respectively, compared with only around 20dB for the rectangular window

(see Fig.17). The rectangular windowed filter has much more passband ripple (see Fig.18).

One negative aspect of the Hamming and Blackman windows is their slower transition from passband to stopband. The Blackman is worse than the Hamming in this respect, but as just noted, provides better stopband attenuation.

The transition bandwidth can be reduced by increasing the number of coefficients. The circuit in Fig.19 includes the same rectangular windowed

filter as in Fig.16, along with a 129-coefficient Blackman windowed filter (also 5kHz cutoff at a 48kHz sampling rate). The 129-coefficient filter was added to the existing .lib file and follows the same pattern as the other examples.

The results are shown in Figs.20 & 21 (decibel/log frequency and linear plots, respectively). In this case, the transition of the Blackman filter (cyan trace) is closer to that of the original rectangular filter (green trace) with this version.

The stopband attenuation of this Blackman windowed filter is the same as previously, at around 74dB at the highest peak in the stopband – increasing the number of coefficients does not change this.

We can obtain very good frequency responses using windowed sinc FIR filters, but they require a large number of coefficients to achieve fast cutoffs. Improving the sharpness therefore requires an increase in run time or faster processing for software implementation, or more circuitry for hardware implementation.

Also, the coefficient values must be accurate to give good results and the typically large range in values between different coefficients means that floating point calculations are often needed.

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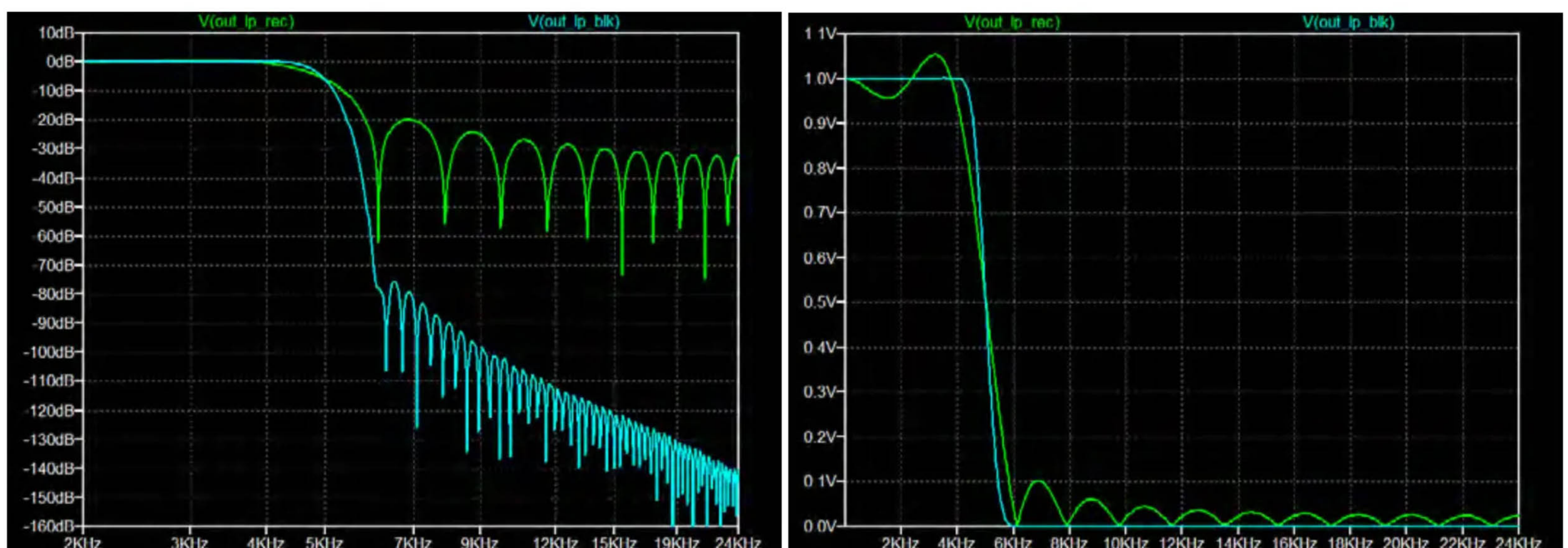


Fig.20: the simulation results from the circuit in Fig.19 (dB/log plot). Fig.21: the simulation results from the circuit in Fig.19 (linear plot).

The Fox Report

Barry Fox's technology column



Going to eleven

It has been said many times before but bears saying again. The real cost of buying a new computer is not the cost of the hardware, but the cost in time and money of getting the new system up and running with all the same software and settings as before.

Programs/apps will need re-installing, which will often only be possible after new activation keys have been bought. Existing peripherals, like printers and scanners and cameras, may or may not work. All the carefully honed settings will need new tweaking. All in all, it's a rotten awful, horribly time-consuming and expensive job.

This is what makes Microsoft's policy of ending support for Windows 10 towards the end of 2025 so nasty. Owners of perfectly good systems will be forced to choose between junking them or risking continued use without the security updates which Microsoft churns out monthly. And you can be sure that criminals will very quickly exploit any security loophole that Microsoft leaves unplugged.

The real devil is in the minimum specification decreed for Windows 11. Whole rafts of PCs have enough memory to clear the upgrade hurdle of 4GB RAM, and they can also meet the Secure Boot and TPM (Trusted Platform Module) security requirements. What fails them is the CPU/processor, for which Microsoft has set a much higher bar.

One of my PCs, a Lenovo desktop, runs Windows 10 very fast and happy with an Intel i7-6700 @3.4GHz. But it flunks the Windows 11 Upgrade test.

The Internet is awash with registry hacks that can be used to fool a Windows 10 PC into ignoring the CPU block. A line of code is added which tells Windows to "AllowUpgradesWithUnsupportedTPMOrCPU".

Adding this line will be easy for anyone familiar with regedit, the Windows registry editing tool. Just Google that phrase. There are also batch files on free download offer, which do the job without the need to use regedit.

But beware trying any of this without first learning how to use a registry editor and making a backup of the existing Registry. For good measure, use the

standard Windows tool to 'Create a Restore Point'.

Furthermore, before trying any forced upgrade, make a full backup of the entire PC, using either the full disk backup software that comes free with Windows or a proprietary backup program such as Macrium (of which sadly there is no longer a free version on offer from the company).

Feeling reasonably confident with regedit and batch files, I added the AllowUpgrades... line to my registry and rebooted to make sure it took effect. But the normal Windows 10-to-11 upgrade procedures were still blocked. Perhaps Microsoft is now doing this intentionally? Their upgrade advice site contains a lot of information on how to buy new equipment. The trade must love all this.

Fortunately, I found it was still possible to force the upgrade by downloading a full ISO copy of Windows 11 from Microsoft, then right-click and Open with Windows Explorer to reveal its contents, which includes a setup.exe file. Click to run this .exe file and the PC goes online to check for updates to the file.

I still got a warning about "not meeting requirements", but clicking the option to Accept ignores the warning.

"Accepting" triggers more checking of files and the PC hardware. Most importantly, you will see a tick box opportunity to "Keep Personal Files and Apps" instead of starting a clean install, which will delete everything! Take no chances here. Double-check you are being offered the "Keep" option and are taking it.

There then follows a long process of more checking, online and locally inside the PC, with disconcertingly lengthy black screens. But, for me at least, it all came right in the end. The PC started again with Windows 10 replaced by Windows 11, and all my old programs, apps, settings and data intact.

To my astonishment, everything worked, except for one error message warning that a driver for an online music site (Idagio) I had since forgotten about had been blocked.

It is possible that other problems will now emerge, like 'blue screens of death' and a refusal to install new updates. So I am keeping my full Windows 10 backup safe in case a delayed action disaster strikes and I have to revert. I am also storing all my new work and data on a separate disk.

Only time will tell whether I have successfully side-stepped Microsoft's grand plan. They seem to want me to junk a tried, tested and nicely tuned hardware and software system and spend heavily on something new that does little more than the same as before.

As a footnote, and purely by way of experiment, I also tried forcing a budget, low-power Acer laptop to upgrade from 10 to 11. Surprisingly, the near-toytown device passed all Microsoft's tests except (unsurprisingly) for the CPU, which is an Intel N3710 that snail-chugs at just 1.6GHz.

After the now-standard registry hack, I got the installation process to start by downloading the Windows 11 ISO direct from the Windows 11 Download page onto the laptop, right-clicking to 'Open With' Windows Explorer and running setup.exe.

The upgrade process began, but it greyed out the vital option to keep all existing apps and data and accused me of "choosing a different language" for the installation than for the laptop settings. I stopped the installation and tried changing the country, language and keyboard – several times in several ways – but the error persisted.

Obviously, losing all data, settings and apps defeats pretty much the whole object of forcing the upgrade. Also, the end result would probably have been an unusably slow laptop. So I gave up on trying to force a Windows 11 upgrade.

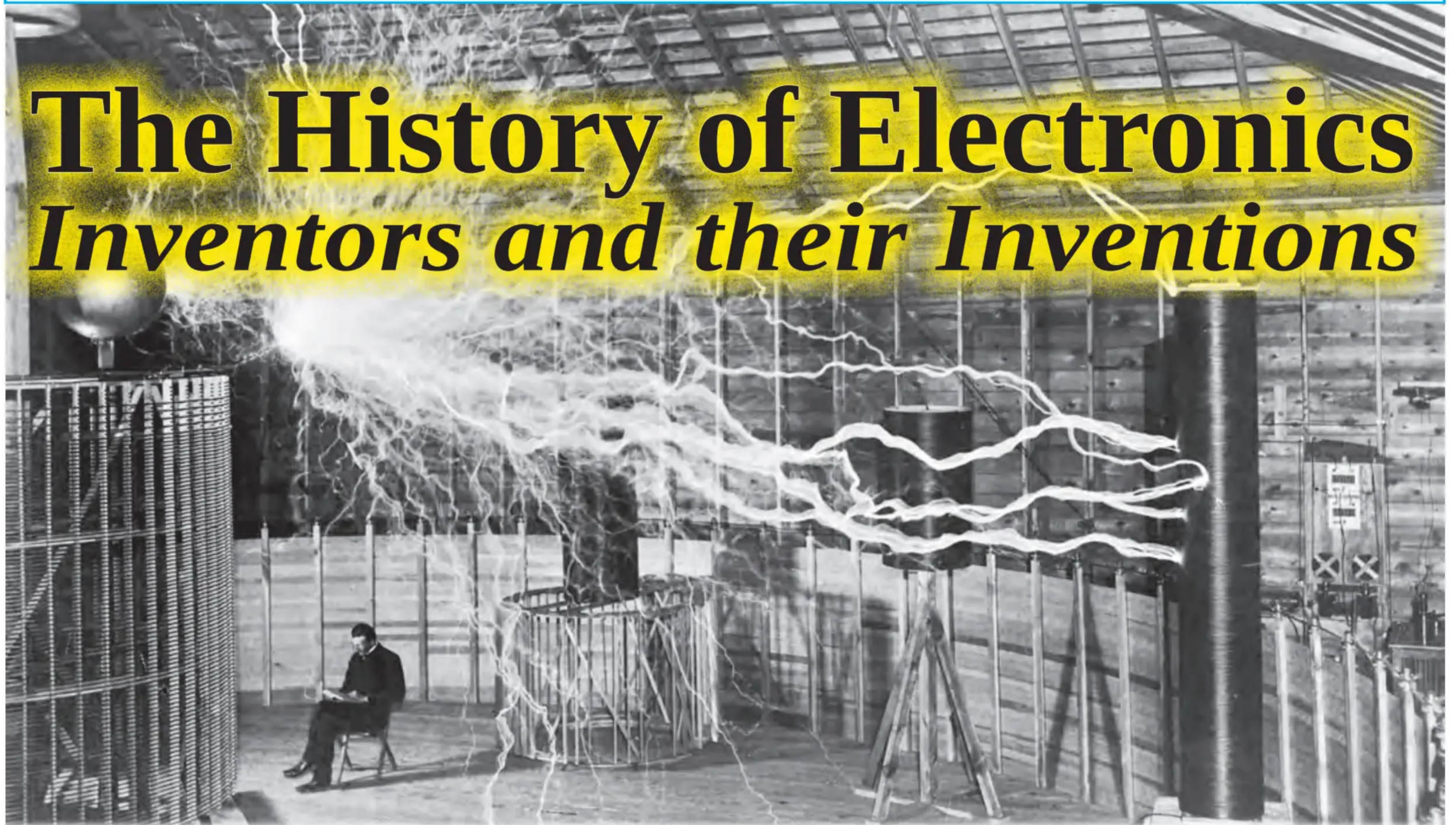
As a further experiment, I then tried to use ChromeOS Flex to turn the Windows 10 laptop into a Chromebook. Downloading ChromeOS was easy but burning it to create the necessary Chrome Recovery USB proved impossible (despite trying several different 8GB and larger USBs).

An Internet search revealed that I am not alone in repeatedly failing and there is no simple one stop fix. So maybe I will end up replacing dead Windows with Ubuntu Linux. Or perhaps I will just let Windows win this one and dump it at my local recycling centre – after first removing or smashing the hard drive, of course.

Editor's note: as an alternative to switching to Windows 11, you might want to consider using the Opatch service mentioned in my Editorial this month. It is said to be able to patch Windows 10 computers for serious security vulnerabilities for at least five years after Microsoft stops releasing official patches.

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The History of Electronics *Inventors and their Inventions*



Physicist Isaac Newton wrote, “If I have seen further it is by standing on the shoulders of Giants”. The field of electronics is no different; we could not have the technology we have today without the contributions of thousands of brilliant people. This series of articles is about them.

Part 3: by Dr David Maddison

The first two articles of this six-part series, published over the last couple of months, listed significant electronics-related inventions of individual inventors born before 1848.

This part will cover individual inventors born from 1848 to 1881, while next month we will continue that list up to the recent past. The last two parts will discuss significant inventions credited to companies and other organisations.

Inventors by date of birth (1848 to 1881):

Shelford Bidwell 1848-1909
photocells

Experimented with selenium photocells in the 1870s, and in 1880, reported how he also duplicated the “photophone” experiment of Alexander Graham Bell (<https://pemag.au/link/abnc>). In another experiment, he used a selenium cell to scan an image and transmit it to another device via wires, which burned the reproduced image onto paper.

He reported the results of his “Tele-Photography” in 1881 (<https://pemag.au/link/abnd>). He was the first to use a photocell to scan an image (Bain and Bakewell did not; see last month). He also invented a device that could scan an original document without redraw-

ing it on special media, analogous to a modern fax machine.

In 1908, he published “Telegraphic Photography and Electric Vision” (<https://pemag.au/link/abne>), on transmitting motion video (ie, TV) and the large amount of data involved.

Chichester A. Bell 1848-1924
tape recorder

Bell and Sumner Tainter (1854-1940) received US patent **341,214** in 1886 for a recording and playback device where sound was recorded on a wax-coated paper strip in a reel-to-reel arrangement.

It was the earliest tape recorder, but was considered inferior to Edison’s wax cylinder for recording and

playback and was not commercially released.

Sir John Ambrose Fleming 1849-1945
thermionic valve, trans-Atlantic transmission

Invented the first thermionic valve, otherwise known as a “vacuum tube”, in 1904 (see Fig.29). Fleming called them oscillation valves “for the rectification of high-frequency electric oscillations as used in wireless telegraphy”. They were diodes, the simplest type of valve.

Fleming’s valve is considered the beginning of electronics because it was the first active electronic component. As radio detectors, Fleming’s valves were not more sensitive than crystal detectors. However, they did not need

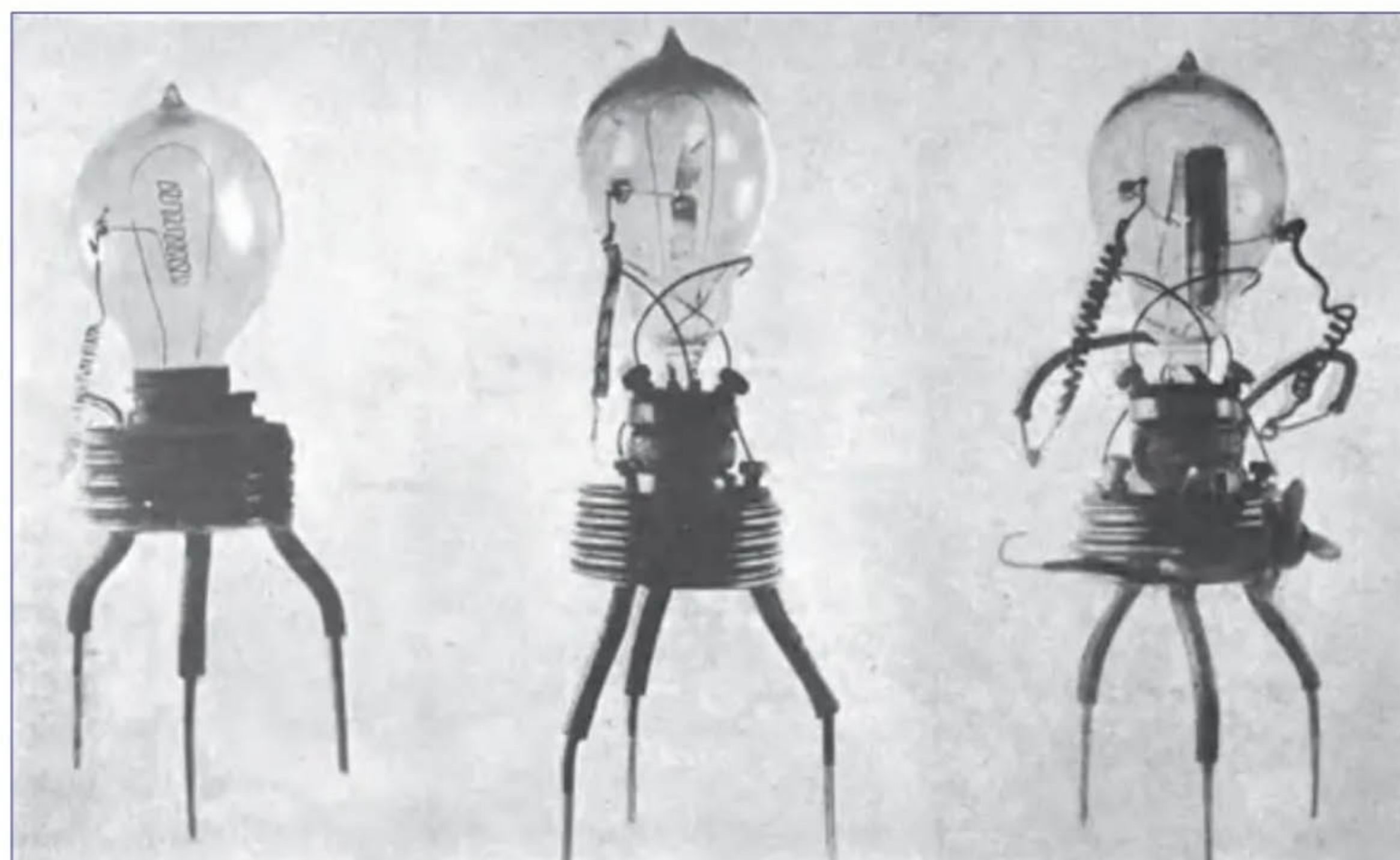


Fig.29: Fleming’s first vacuum tube diodes from 1904. Source: <https://w.wiki/7DAU>

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constant adjustment for use on ships due to the movement like crystals did.

In 1899, Fleming, under contract from Marconi, designed the first high-power radio transmitter, much larger than the 200-400W transmitters used by Marconi. It was a spark-gap transmitter powered by a 25kW alternator and it performed the first trans-Atlantic transmission in 1901, over 3500km, which was credited to Marconi despite Fleming's involvement.

Charles Fritts

1850-1903

solid-state solar cells

He made the world's first solid-state solar cells in 1883 with selenium and a thin layer of gold. They had an efficiency of 1%, making them too expensive and inefficient for generating power, but they were used as light sensors for cameras and in other applications into the 1960s.

Oliver Heaviside

1850-1925

mathematical equations, E region etc

Reformulated and simplified Maxwell's equations to make the **Maxwell-Heaviside equations** and put them in their modern form.

He also invented the **Heaviside step function** to calculate the current drawn when an electrical circuit is switched on, and developed transmission line theory (or telegraphers' equations). The latter increased the transmission rate of the trans-Atlantic telegraph cable ten times, to one character per minute.

He discovered that telephone line transmissions could be improved by a series inductance in the cable. He and Arthur Edwin Kennelly (1861-1939) independently predicted the presence of the Heaviside layer, Kennelly-Heaviside layer or E region, part of the ionosphere that reflects medium-frequency waves.

William Edward Sawyer

1850-1883

electricity distribution

Sawyer worked on telegraphy and electric lighting. With Albon Man, he

founded the Electro-Dynamic Light Company (1878-1882), later purchased by Westinghouse, to provide lighting and distribute electricity into cities. His lighting system contained a safety switch and a current regulator.

His company had patents dated 1877 and 1878 for incandescent lights, predating Edison. Sawyer's lights were not long-lasting, a problem he never solved.

Karl Ferdinand Braun

1850-1918

cathode ray tube (CRT), oscilloscope etc

He discovered the rectifying properties of a metal-semiconductor junction (schottky diode) in 1874, using mercury as the metal and copper sulfide or iron sulfide as the semiconductor.

He also invented the cathode ray tube and the oscilloscope in 1897. He worked on wireless telegraphy and invented a crystal detector in 1898, among other contributions. In 1905, he devised the phased array antenna.

Edward Weston

1850-1936

Weston Cell, Constantan & Manganin alloys

Invented the Weston Cell in 1893, a highly-stable electrochemical cell used as a voltage reference. It was the international standard for EMF from 1911-1990.

He invented the alloy Constantan in 1887, which has a low variation in resistivity with temperature, used in thermocouples, and Manganin in 1892, with almost no variation in resistivity with temperature, used in precision resistors.

In 1888, he founded the Weston Electrical Instrument Corporation, which became famous for the wide variety of high-quality electric meters it manufactured.

Sir Oliver Joseph Lodge

1851-1940

moving-coil loudspeaker etc

Lodge identified electromagnetic radiation independent of Hertz. He also made an improved Hertzian wave detector based on metal filings in a tube he

named a "coherer", based on Branly's earlier work (see last month).

Under the influence of a radio signal, the conductivity between the two electrodes would change. The device had to be regularly tapped to restore its sensitivity. It was used until 1907, when Marconi's crystal detector replaced it. He also invented the moving-coil loudspeaker in 1898.

In 1898, he invented and patented "syntonic tuning" to tune radio equipment to specific frequencies, causing a patent dispute with Marconi. He developed a form of electric spark ignition for internal combustion engines.

Emile Berliner

1851-1929

microphone, Berliner Gramophone Record

Developed an improved type of telephone transmitter (microphone); his patent was acquired by the Bell Telephone Company. It was contested by Thomas Edison, who won the case. There were many expensive and complicated court cases in the USA in the 1870s and 1880s contesting the invention of the telephone; see <https://www.wiki/7DYI>

In 1887 and 1888, Berliner received US patents [372,786](#) & [382,790](#) for the "Berliner Gramophone Record". They were flat discs, like the records we know today, although the Berliner records were only 18cm in diameter, played two minutes per side and rotated between 60RPM and 75RPM. They competed against wax cylinder recordings. There is a project to put about 18,000 Berliner recordings on Flickr: <https://pemag.au/link/abpa>

Leonardo Torres y Quevedo

1852-1936

"Telekino" remote control, El Ajedrecista game

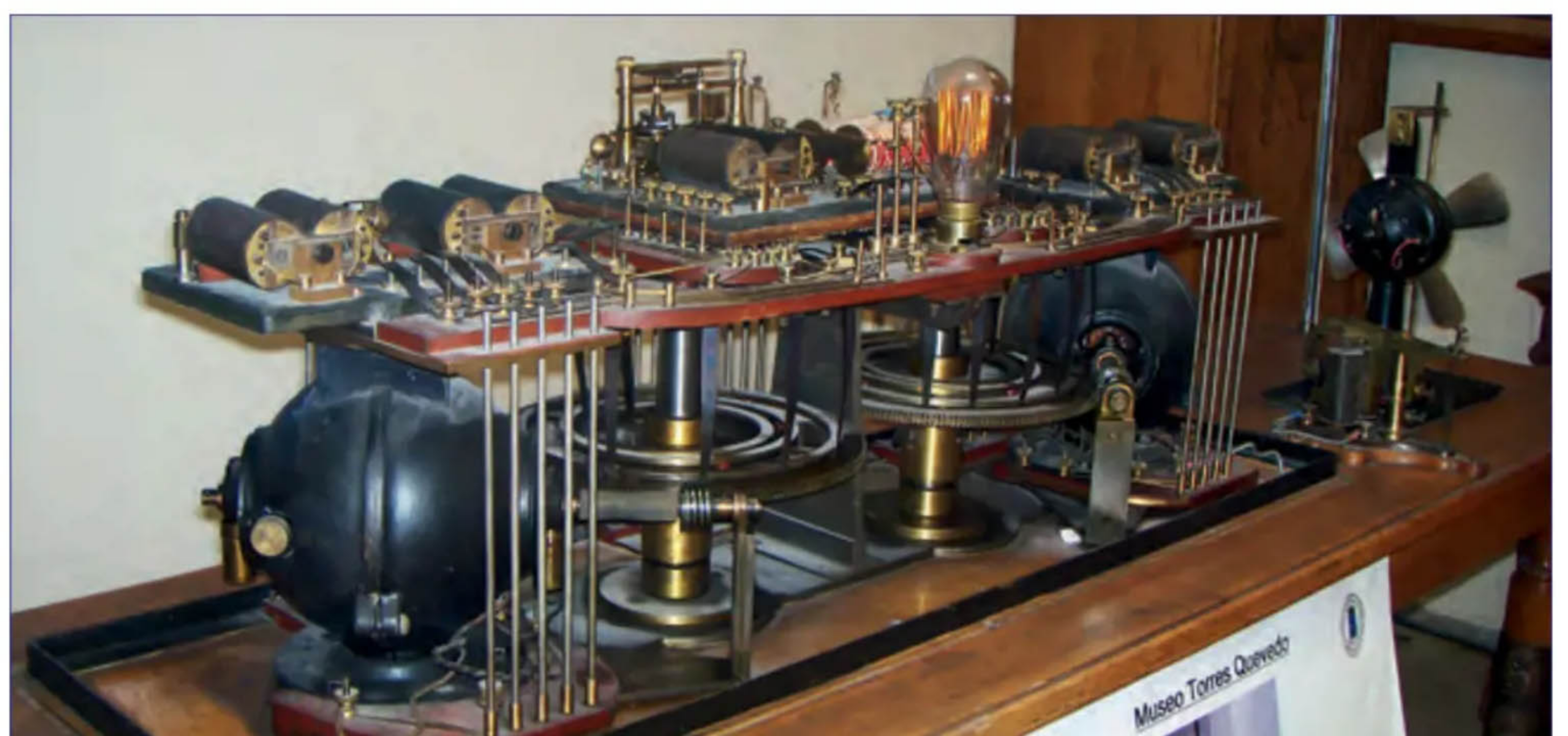
Quevedo demonstrated a remote control he invented in 1903, called the "Telekino" (Fig.30). It was remarkably advanced for the time and was the second remote control invented after Tesla's in 1898. 19 different commands could be sent, with the command



Karl Ferdinand Braun was a founder of Telefunken.

Source: www.cathodique.net/FBraun.jpg

Fig.30 (right): the Telekino receiver in the Torres Quevedo Museum in Madrid, Spain.
Source: <https://www.wiki/7DAV>



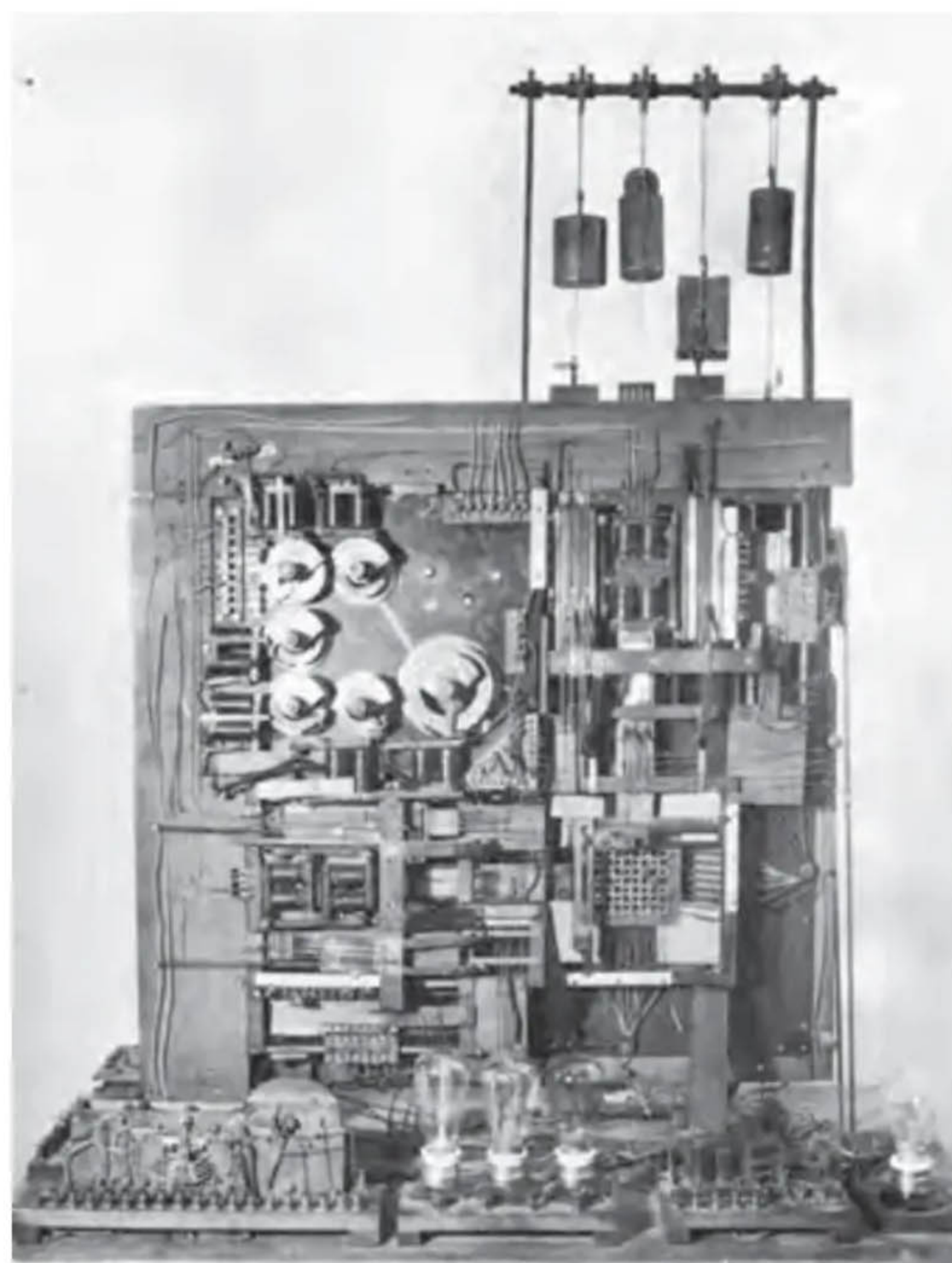


Fig.31: the remarkable El Ajedrecista chess-playing machine. Source: www.torresquevedo.org/LTQ10/images/PrimerAjedrecista.jpg (CC BY-SA 3.0).

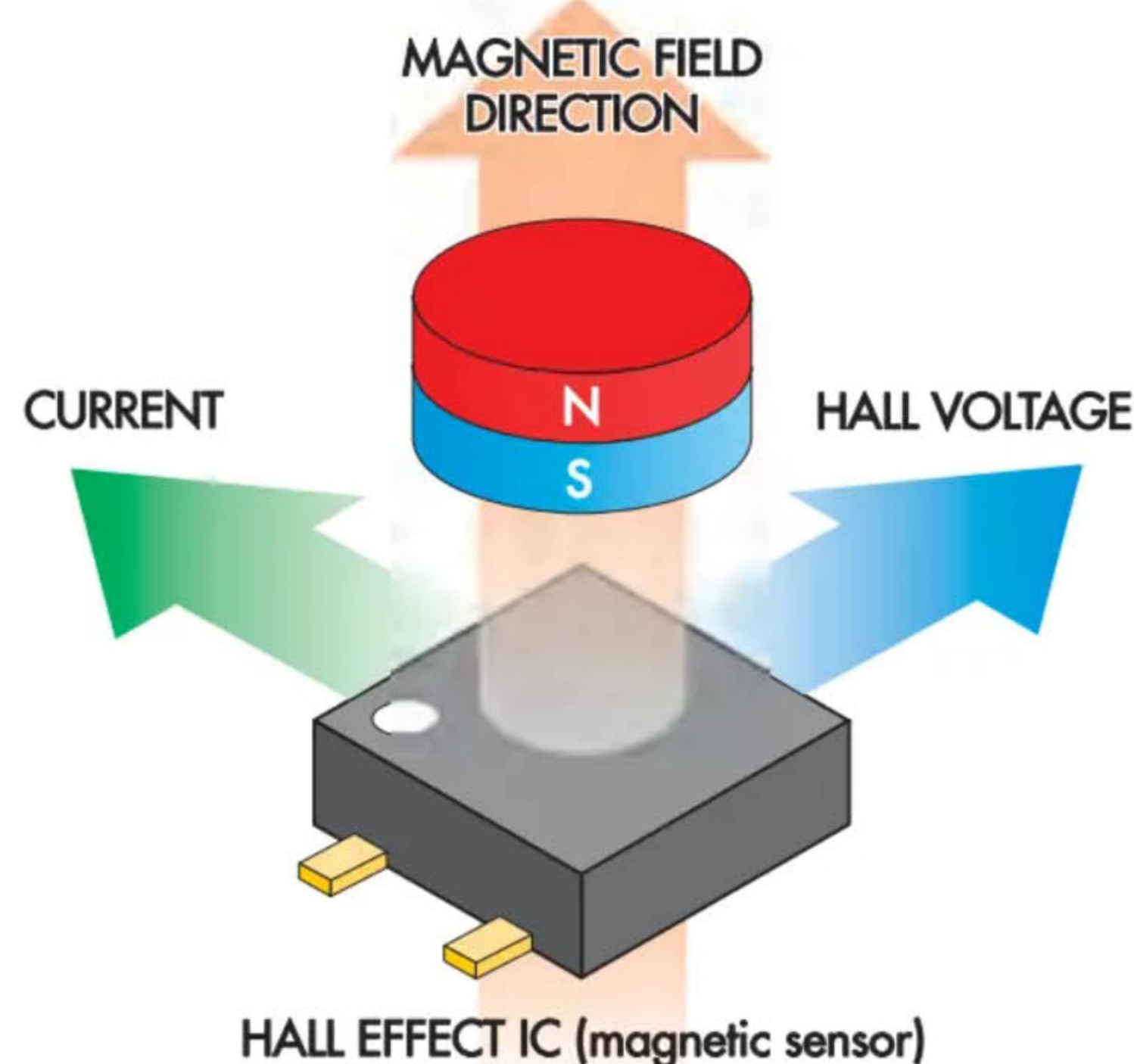


Fig.32: the operation of a Hall effect IC. Original source: www.ablic.com/en/semicon/products/sensor/magnetism-sensor-ic/intro/

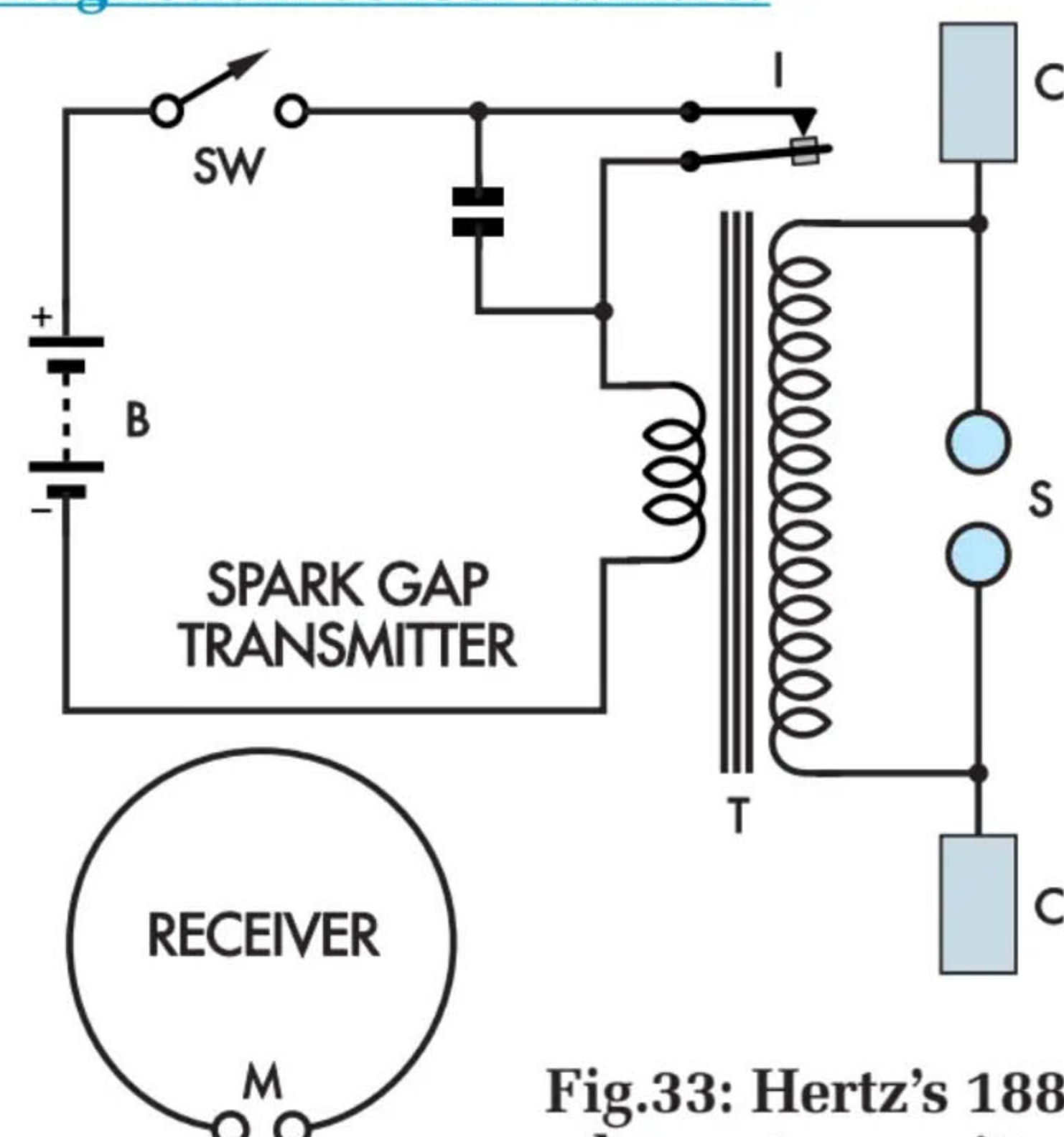


Fig.33: Hertz's 1887 spark-gap transmitter, with an induction coil, dipole antenna, capacitance (C) at the ends, a spark gap (S) and resonant loop antenna receiver with a spark micrometer (M) to measure signal strength. It operated at around 50MHz. Source: <https://wiki/7DAW> (CC-BY-SA-3.0).

sequence recorded. He tested it with dirigibles in 1901.

In 1905, he demonstrated the device with a three-wheeled vehicle, and in 1906, a boat with people onboard. The work was abandoned due to a lack of money. He also invented what was arguably the first computer game. It was called "El Ajedrecista" and could play certain chess moves (see Fig.31).

Mechanical arms moved pieces while sensors detected the opponent's moves. It still works today and can be seen at the Torres Quevedo Museum in Madrid.

Temistocle Calzecchi-Onesti 1853-1922
experiments leading to the coherer

Conducted experiments from 1884 on the electrical conductivity of tubes of metal filings and how they were affected by various electrical influences. This led to Branly's invention of the coherer (see Lodge's entry on page 53).

Heike Kamerlingh Onnes 1853-1926
superconductivity

He discovered superconductivity in 1911 (the loss of all electrical resistance of some materials at certain low temperatures). It is used to generate powerful magnetic fields in machines like MRI scanners. High-temperature superconductors with less stringent cooling requirements are currently being developed.

Jonas Wenström 1855-1893
three-phase electrical system

Received a Swedish patent for a three-phase electrical system in 1890. He developed it independently of Mikhail Dolivo-Dobrovolsky (see his entry on page 56).

Edwin Herbert Hall 1855-1938
Hall effect

He discovered what is now known as the **Hall effect** in 1879, the basis of modern magnetic field detectors and Hall thrusters on spacecraft. It explains that a voltage is produced at right angles to a current flow in a conductor with a magnetic field perpendicular to the current flow – see Fig.32.

Paul-Jacques Curie 1855-1941
piezoelectricity

With his brother Pierre Curie (1859-1906), discovered piezoelectricity (used for guitar pickups etc) in 1880. They also studied pyroelectricity.

Nikola Tesla 1856-1943
polyphase electrical system, Tesla coil etc

Tesla was a prolific inventor and genius. He developed the polyphase electrical system (AC power with more

than one phase) and associated induction motors, licensed by Westinghouse in 1888.

From 1890, he tried to develop a wireless lighting system using Geissler tubes powered by a Tesla coil he invented in 1891.

He was photographed at his Colorado Springs facility in 1899 with the "magnifying transmitter" Tesla coil (done using double-exposure; see the lead image). It produced 12MV 150kHz arcs up to 41m long with an input power of 300kW.

In 1893, he consulted on the design of a Niagara Falls hydroelectric power station. In 1898, he developed the first wireless radio remote control for a boat, a concept he called teleautomatics. In 1906, he demonstrated a bladeless turbine for a power station, which spun at 16,000RPM and produced 150kW.

The unit of magnetic flux intensity, the **tesla (T)**, is named after him.

Sir Joseph John Thomson 1856-1940
acoustic waveguide

Contributed to atomic physics. In 1893, he proposed the acoustic waveguide, and in 1894, Oliver Lodge experimentally verified it. In 1897, Thomson suggested the existence of the electron. He also conducted experiments with cathode rays.

Heinrich Rudolph Hertz 1857-1894
spark gap transmitter, radio waves

Hertz proved the existence of radio waves, first predicted by Maxwell's equations, from 1887 onward. He demonstrated properties such as polarisation, reflection and standing waves. In 1887, he also built the first spark gap transmitter (Fig.33). The unit of frequency, the **hertz (Hz)**, is named after him.

William Stanley Jr 1858-1916
AC transformer and complete AC system

Built the first practical AC transformer in 1885 based on the prototype of Gibbs and Gaulard; see US patent [349,611](#). In 1886, he demonstrated a complete AC system with generators, transformers and high-voltage transmission lines in Great Barrington, Massachusetts, lighting offices and stores.

Sir Jagadish Chandra Bose 1858-1937
millimetre waves, microwave components etc

He produced millimetre (5mm wavelength) 60GHz electromagnetic waves in 1894 because they were a more convenient size to work with in his small laboratory – see Fig.34.

In 1895, he showed that millimetre

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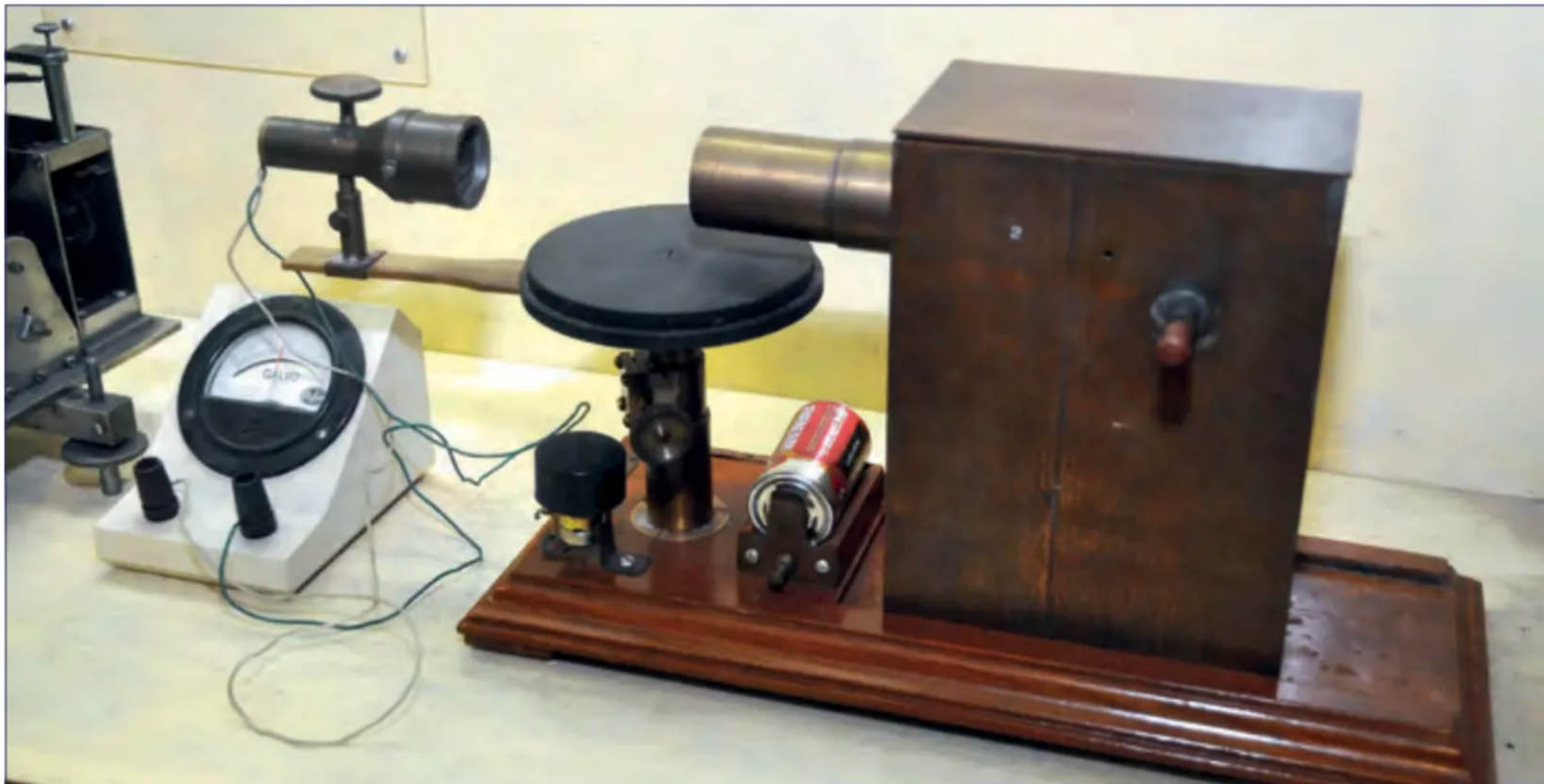


Fig.34: 60GHz microwave apparatus by Jagadish Bose. The galvanometer and battery are modern. The transmitter on the right generates microwaves from sparks between tiny metal balls. Above the galvanometer is a galena point-contact detector inside a horn antenna. Source: <https://w.wiki/7DAY> (CC-SA-3.0).

waves could go through the human body and walls, achieving a range of 23m. Bose was not interested in patenting or commercialising his amazing work, although he was persuaded to patent a metal-semiconductor diode in 1901, awarded in 1904 (US patent [755,840](#)).

He developed a galena semiconductor crystal microwave detector and many other now-familiar microwave components, such as waveguides, horn antennas, dielectric lenses and polarisers. Much of his equipment can be seen at the Bose Institute Museum in Kolkata, India (www.jcbose.ac.in/museum).

Nobel laureate Sir Neville Mott said that Bose was 60 years ahead of his time and that he had antici-

pated p-type and n-type semiconductors. One of his concepts from a paper he wrote in 1897 was used in the 1.3mm multibeam receiver of the National Radio Astronomy Observatory (NRAO) 12m telescope in Tuscon, Arizona.

Friedrich August Haselwander 1859-1932
electric arc lamp

Invented an electric arc lamp in 1880, and in 1887 invented and put into service a synchronous three-phase generator in Europe (Fig.35). It developed about 2.8kW at 960RPM and 32Hz.

Aleksandr Popov 1859-1906
lightning detector (radio receiver)

Popov built a wireless lightning detector in 1895 (see Fig.36), one of the first radio receivers, and in 1896



Fig.35: Haselwander's three-phase generator with stationary ring armature and four-pole rotor, as displayed in 1891 at the International Electrotechnical Exhibition in Frankfurt. Source: <https://w.wiki/7DAZ>

transmitted radio signals over 250m. Some of his work was based on the findings of Sir Oliver Joseph Lodge.

In 1898, he performed ship-to-shore communication using wireless telegraphy over 10km, and in 1899, 48km. In ex-USSR countries, the 7th of May is celebrated as Radio Day, the day Popov first demonstrated his lightning detector.

Herman Hollerith 1860-1929
punch(ed) cards

Developed punched cards for data storage and analysis, used in the 1890 US Census. These evolved into IBM punched cards, used as late as the early 1980s. See SILICON CHIP magazine's January 2023 article on Computer Memory for more on punched/punch cards (siliconchip.au/Series/393).

Ottó Titusz Bláthy 1860-1939
modern transformers, voltage regulator etc

Bláthy, Károly Zipernowsky (1853-1942) and Miksa Déri (1854-1938) applied for a patent for the first modern transformers in 1885, which were much more efficient than the designs of Gaulard or Gibbs. The trio also designed the first power station with AC generators "to power a parallel-connected common electrical network".

Bláthy also invented the voltage regulator, AC watt-hour meter (1889), motor capacitor for single-phase AC motors and turbo generator for steam power plants.

Paul Julius Gottlieb Nipkow 1860-1940
Nipkow disc

Invented the Nipkow disc in 1883. It was a disc with a spiral pattern of holes to divide a picture into a linear series of points to enable opto-electronic



Sir Jagadish Bose demonstrating the horn antenna. Source: <https://w.wiki/7DuL>

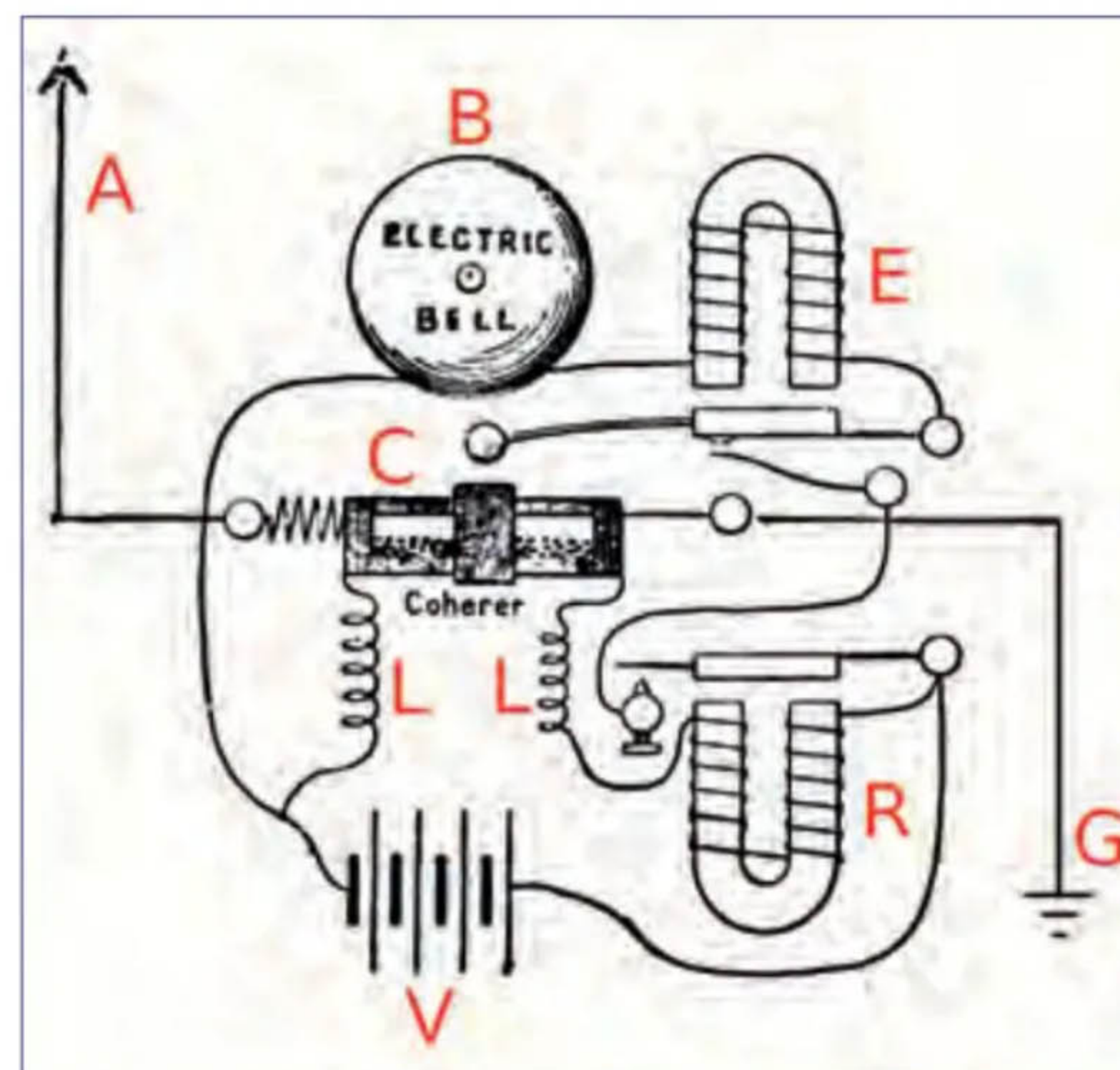


Fig.36: Alexander Stepanovich Popov's 1895 "coherer receiver", one of the first radio receivers, designed to detect lightning strikes. Key: A) antenna, B) bell, C) coherer (detector), E) electromagnet, G) ground, L) chokes for noise immunity, R) relay, V) battery. Source: <https://w.wiki/7DAa>

Feature Article



Paul Gottlieb Nipkow is considered to be one of the fathers of television. Source: <https://w.wiki/7DuZ>

imaging of an object. There was little interest at the time.

It became the basis of the first electro-optical television systems in the 1920s-30s (see SILICON CHIP magazine's articles on Display Technologies in the September & October 2022 issues for more - siliconchip.au/Series/387).

Peter Cooper Hewitt 1861-1921
mercury vapour lamp, mercury arc rectifier

He invented the mercury vapour lamp in 1901, the predecessor of the fluorescent lamp. In 1902, he invented the mercury arc rectifier, the first commercially available non-mechanical rectifier. In 1916, he was involved in developing the Hewitt-Sperry Automatic Airplane, the predecessor of the cruise missile.

Mikhail Dolivo-Dobrovolsky 1862-1919
asynchronous three-phase motor

Invented an asynchronous three-phase motor in 1888, which had low torque at low speeds. This problem was solved with a variation of that motor, the slip-ring motor, with high torque at low speeds in 1891. He also developed the delta-wye transformer for three-phase distribution systems in that year.

Philipp Lenard 1862-1947
Lenard window (aluminium) for Crookes tube

Lenard began investigations of cathode rays in 1888 and developed a modified Crookes tube with what was to become known as a "Lenard window", a thin aluminium window that made it possible to study the radiation from outside the tube.

Alan Campbell-Swinton 1863-1930
transmitting television images

Alan Archibald Campbell-Swinton experimented with cathode ray tubes (CRTs) in 1903 for transmitting television images.

Prompted by Shelford Bidwell, on the 18th of June, 1908, his letter in Nature entitled "Distant Electric Vision" (<https://pemag.au/link/abpb>) said the problems of mechanical television

would be solved by electronic systems with CRTs at both ends.

Walther Hermann Nernst 1864-1941
Nernst (incandescent) lamp

Invented the Nernst lamp (Fig.37) in 1897 as an improvement to the incandescent lamp. The way it works is very interesting. An element heats a ceramic rod made of zirconium oxide and yttrium oxide. The rod's resistance decreases as it heats up and the heating element is turned off. A current sustains the glowing ceramic rod due to ohmic heating.

It can operate in the air, as the ceramic rod will not degrade like a metal filament. They are obsolete as a visible light source but are still used as an infrared light source in spectroscopy, as they emit infrared over a wide range of wavelengths. See the video titled "The Nernst Lamp" at <https://youtu.be/1vCQySb6uLA>

Charles Proteus Steinmetz 1865-1923
Steinmetz's equation

He contributed to AC hysteresis theory from 1890 and solved practical problems with heat build-up in AC motors. This resulted in him building a powerful motor for Otis Elevators to reach higher floors.

His work led to Steinmetz's equation for calculating losses in magnetic core materials, published in 1892 (see the PDF at <https://pemag.au/link/abnf>).

Parts of the Nernst Lamp

The elements of the Nernst Lamp are the glower, heater (made up of two or four heater tubes), ballast and cut-out. These are assembled in the lamp body and the holder.

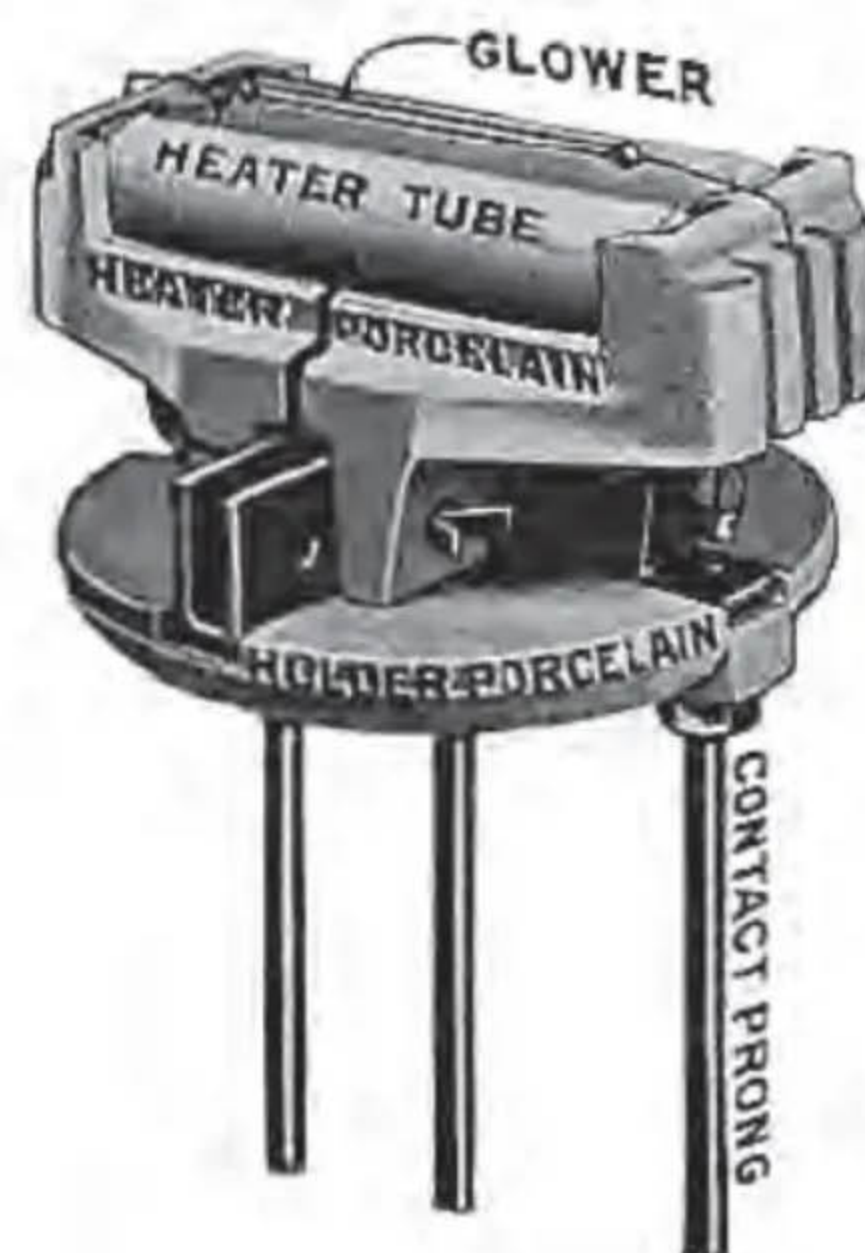


FIG. 3. NAMES OF PARTS OF THE NERNST LAMP HOLDER

Glower The glower, or light giving element, is a white porcelain-like rod about $\frac{1}{2}$ inch in diameter by 1 inch long. It is fastened to the holder mechanically and electrically by means of terminal wires and small aluminum plugs.

Fig.37: a Nernst lamp, an early form of incandescent light. Source: <https://w.wiki/7DAb>

He worked on AC circuit theory and analysis, which he greatly simplified from previous methods, announcing his findings in 1893. He also investigated AC transient theory and other transient phenomena, such as lightning bolts.

Reginald Aubrey Fessenden 1866-1932
radio, sonar etc

He first transmitted speech by radio in 1900 and made the first two-way radiotelegraphic communication across the Atlantic in 1906. He invented an electroacoustic transducer called the Fessenden oscillator in 1912, and in 1914, it detected icebergs 3km away. It was also used for underwater telephony and depth sounding.

For more information on that, see SILICON CHIP magazine's June 2019 article on Bathymetry (siliconchip.au/Article/11664).

Marie Curie 1867-1934
mobile X-ray machine

Invented the mobile X-ray machine in around 1915, powered by a dynamo.

Henri Abraham 1868-1943
astable multivibrator

He and Eugene Bloch (1878-1944) invented the astable multivibrator. The work was done during WW1 but published in 1919. He made the first measurements of the speed of electromagnetic propagation between 1911-1914 and developed the first French triode valve.

John Stone Stone 1869-1943
telephone links, selective four-circuit tuning

Worked in telephone technology and radios. In 1890, he started work on a mathematical analysis of telephone links for American Bell Telephone Co. In 1900, he developed "selective four-circuit tuning" for radios to improve their selectivity and reduce noise.

Boris Lvovich Rosing 1869-1933
early television

He started considering ideas of what we now know as television in 1897, but he called it the "electric telescope". His approach for the receiver was purely electronic, using a CRT, unlike other ideas for television around that time that were mainly mechanical.

By 1902, he made a device that could draw a basic figure on a CRT. Instead of a slow-reacting selenium cell detecting light for the camera, he used a fast-reacting photocell onto which the image was projected by a rotating mirror system.

He obtained patents for his invention in 1907 and 1911. It was presented

The History of Electronics, part 3

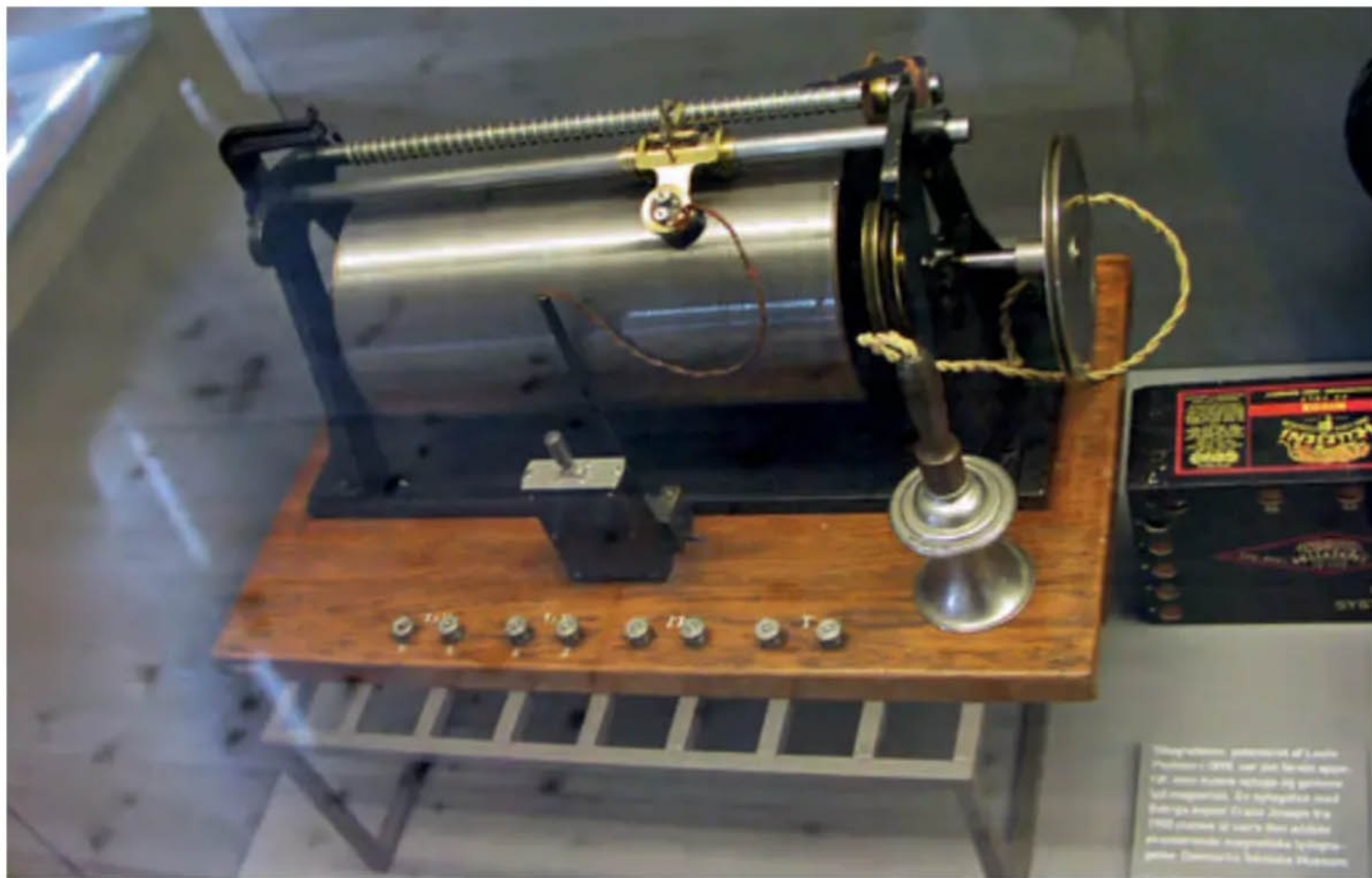


Fig.38: Valdemar Poulsen's magnetic wire recorder, invented in 1898. Source: <https://w.wiki/7DA4> (CC-BY-SA-2.5).

in *Scientific American*, 1st of April, 1911:

<https://pemag.au/link/abng>

Valdemar Poulsen 1869-1942
magnetic audio recordings

Successfully implemented the first means to magnetically record audio in 1898 by magnetising wire along its length (Fig.38). There was no amplification, so the recording was faint but audible with headphones. The device was called the Telegraphone and had limited commercial success due to its low volume and complexity.

With his assistant, Peder Oluf Pedersen, he developed other recording devices using tape and discs. In 1903, he also invented the Poulsen Arc Transmitter (Fig.39), widely used as a radio transmitter in the early 1920s before vacuum tubes were developed.

Arthur Korn 1870-1945
fax machines

Korn pioneered the modern fax machine, which he used to transmit photographs. He used light-sensitive selenium cells in his "phototelauto-graph" or "Bildetelegraph". In 1906, he sent a photo of Crown Prince Wilhelm over 1800km via the telegraph network.

In 1913, he transmitted a movie recording, although the specifics are unclear. We assume it was a frame-by-frame transmission. In 1923, German police used Korn's system to transmit photos and fingerprints.

Paul Langevin 1872-1946
ultrasound transducer

Invented the first ultrasound device in 1917, the quartz sandwich transducer for submarine detection.

William David Coolidge 1873-1975
making tungsten ductile

Coolidge developed a method for

making tungsten ductile, allowing it to be drawn into filaments for light globes. The globes were sold by General Electric from 1911.

Lee De Forest 1873-1961
three-element triode, recording audio

Invented a three-element triode thermionic tube in 1906, the "grid Audion" (Fig.40), for use as an amplifier and an oscillator. This invention is regarded as the start of the Electronic Age.

In 1919, he patented the DeForest Phonofilm system for optically recording audio waveforms onto movie films.

Guglielmo Marconi 1874-1937
wireless transmission, spark gap transmitter etc

Guglielmo Giovanni Maria Marconi built a device to receive radio waves produced by lightning in 1894. That year, he also demonstrated wireless transmission to ring a bell across a room.

He developed a spark gap transmitter and coherer receiver. A coherer was a glass tube with metal filings that radio waves caused to become closer together and therefore more conductive.

In 1895, he designed a system that could transmit over 3km. By 1896, he had transmitted over 6km, then 16km. In 1899, he transmitted across the English Channel.

In 1907, he established a commercial trans-Atlantic telegraph service.

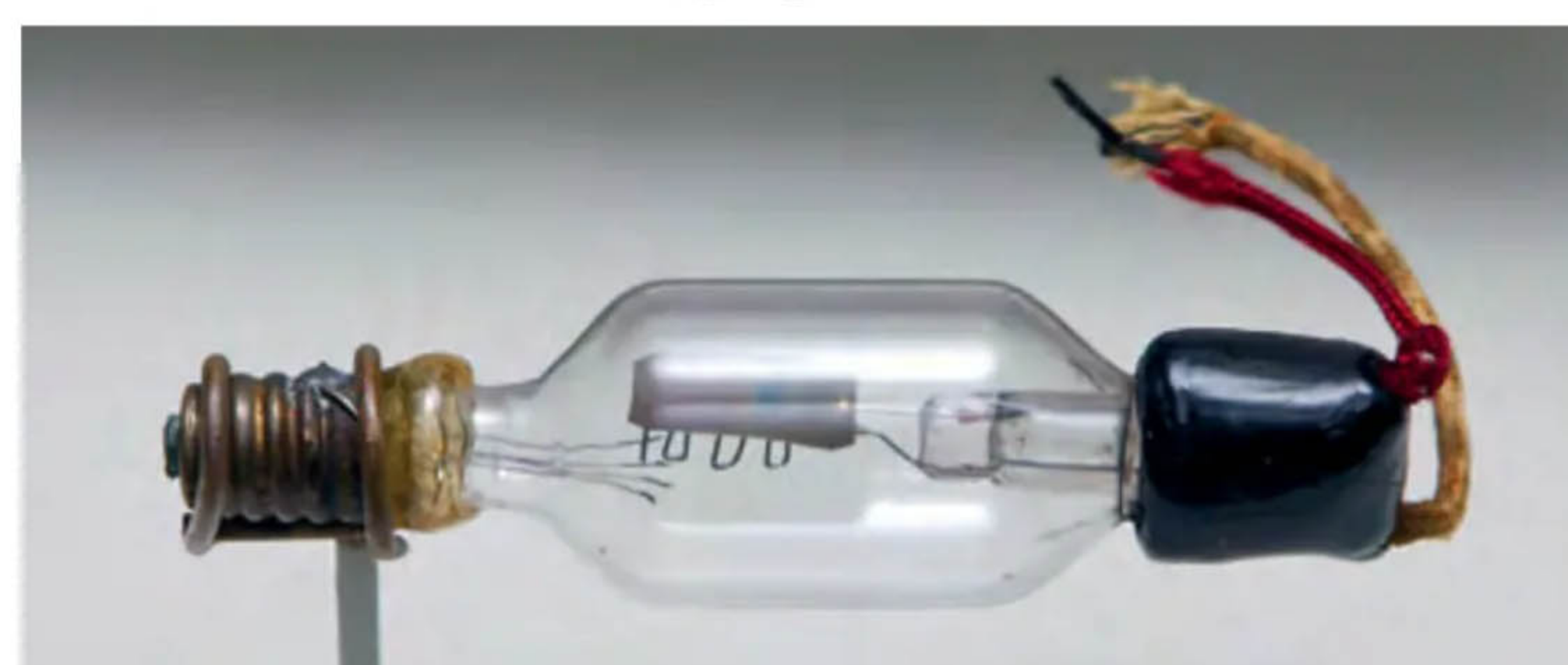


Fig.40: an early grid Audion, invented in 1906, which many regard as indicating the start of the electronic age. Source: <https://w.wiki/7DAf> (GFDL-1.2).

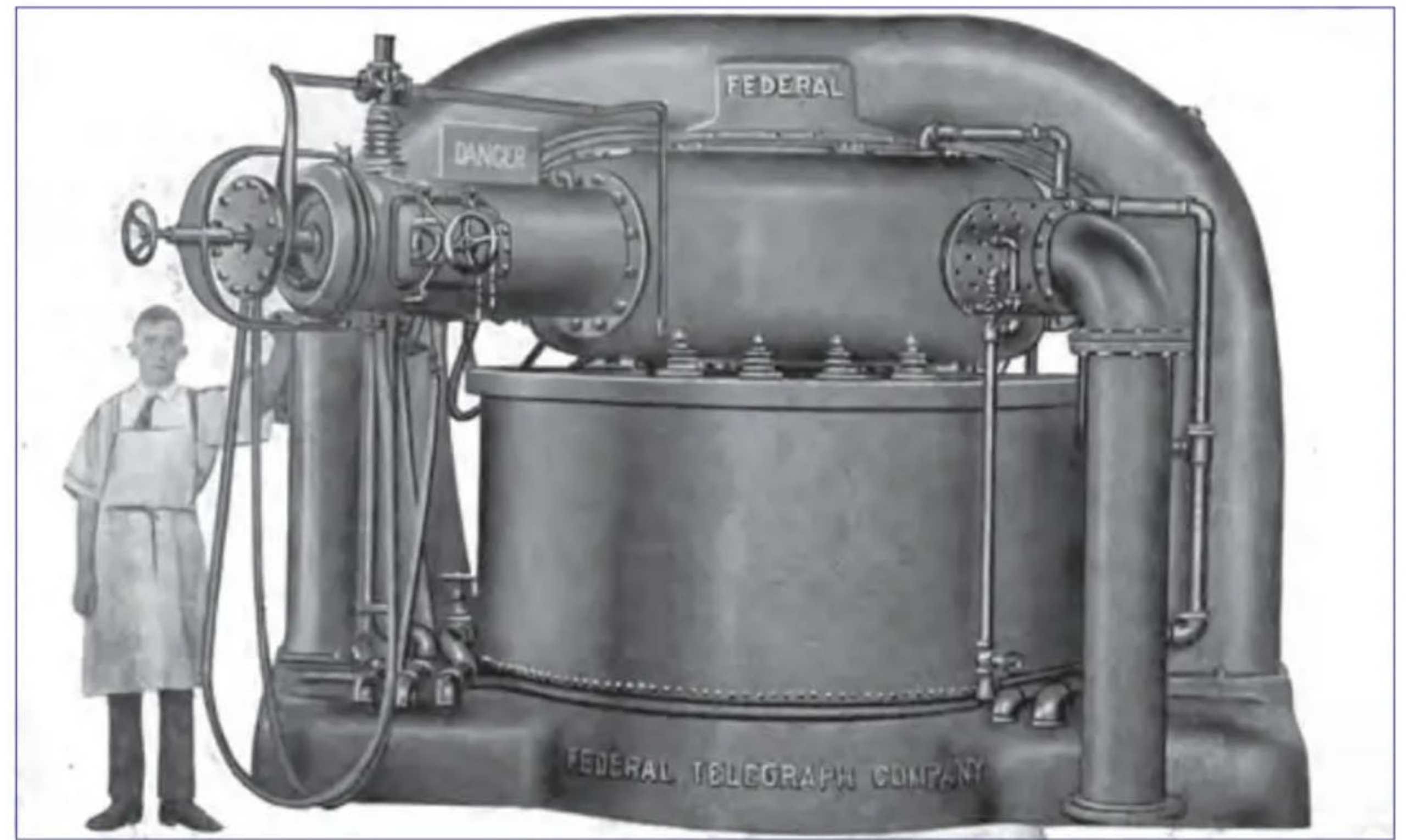


Fig.39: a 1919 Poulsen arc transmitter from a US Navy radio station with a continuous power rating of 500kW (1MW short-term). Source: <https://w.wiki/7DAe>

Marconi's radios were important in rescuing survivors of the RMS Titanic (1912) and RMS Lusitania (1915).

Just Sándor Frigyes 1874-1937
tungsten filament light globe

Also known as Alexander Friedrich Just, he and Franjo Hanaman (1878-1941) were the first to invent an incandescent light globe with a tungsten filament in 1904. They were brittle due to the way they were made, although they lasted longer and were very efficient compared to carbon filaments.

They licensed their patent to the Tungsram company (which also licensed Bródy's patent for using krypton gas in globes in 1934). In 1904, they applied for a Hungarian patent and, in 1905, applied for US Patent [1,018,502](#). The tungsten filament globe became practical with the invention of Coolidge's fabrication method for tungsten filaments.

Alexander M. Nicholson unknown
crystal oscillator

Invented the first crystal oscillator in 1917 using Rochelle salt, a piezoelectric material, and filed for US patent [2,212,845](#) in 1918.

Walter Guyton Cady 1874-1974
quartz crystal oscillator

He invented a quartz crystal oscillator in 1921 and realised that such devices could be used as frequency standards. He filed US patents in 1921 ([1,472,583](#)) and 1937 ([2,170,318](#)).

Feature Article

Édouard Belin 1876-1963
Bélinographe, image transmission

Invented the Bélinographe, which used a photocell to scan and transfer press photos (see Fig.41). It was developed in 1907 and first used commercially in 1913-1914 to transmit pictures over dedicated leased telephone lines. Later models could use ordinary telephone lines.

In 1921, a version was used to transmit a photo by radio across the Atlantic. By 1926, RCA was using it to transmit Radiophotos.

Miller Reese Hutchison 1876-1944
electronic hearing aid, tachometer, Klaxon

He developed the first commercial electronic hearing aid in 1898 (going to market in 1899), with a carbon microphone he called the "Akoulallion". In 1900, he developed a portable battery-powered device that he called the "Akouphone", then the Acousticon 1 in 1902. For more details, see <https://pemag.au/link/abnl>

In 1908, Hutchison invented an electric tachometer for ships and the Klaxon in 1910.

Robert Von Lieben 1878-1913
triode with control grid

Lieben, with engineers Eugen Reisz and Siegmund Strauss, invented the gas-filled (low vacuum) triode with a control grid in 1910. It was the first thermionic valve designed for amplification rather than demodulation and was used as a telephone repeater.

Ernst Alexanderson 1878-1975
Alexanderson alternator

Ernst Frederick Werner Alexanderson invented the Alexanderson alter-

nator in 1904. It produced radio waves more efficiently and with a narrower bandwidth than the spark-gap transmitters used until that time, and it could deliver them continuously at high power.

Alexanderson alternators were used to transmit long-wave radio communications from shore stations from 1906 to the 1990s, although they were too big and heavy for most ships.

The first commercial model would generate a frequency of 100kHz and had a power rating of 50kW. The last transmitter in regular use was Grimeton Radio Station in Sweden, which was used until 1996 and is occasionally used today (see page 17 of SILICON CHIP's March 2023 issue).

Albert Einstein 1879-1955
theory of relativity, photoelectric effect

Published his theory of relativity in 1905. Relativity must be considered in operating satellite navigation systems such as GPS and many other applications.

He also explained the photoelectric effect in 1905, expanding on the work of Planck, which went on to be used in night-vision devices, among others.

Alexander Behm 1880-1952
echo sounding, Echolot

He invented echo sounding in 1912 to measure water depth and detect obstacles, obtaining a patent in 1913. In 1922, he produced the Echolot to measure water depth beneath a ship.

Albert W. Hull 1880-1966
dynatron vacuum tube, magnetron

Invented the dynatron vacuum tube in 1918 and the magnetron in 1920,

which was used as an amplifier and low-frequency oscillator. The latter is still used in microwave ovens (albeit in a modified form; see the entry for Russell Harrison Varian next month).

Louis Blattner 1881-1935
Blattnerphone

Blattner, under license by Kurt Stille (1873-1957), produced a new audio recorder using steel tape instead of wire called the "Blattnerphone" in 1925. It was also based on the magnetic recorder of Valdemar Poulsen (see his entry on page 57).

In 1933, the Marconi Company acquired the rights to the Blattnerphone and made an improved version called the Marconi-Stille recorder, which the BBC used from 1935 into the 1940s (Fig.42).

Irving Langmuir 1881-1957
vacuum pump improvements

Improved the vacuum pump, which led to high-vacuum rectifiers and amplifier tubes. He, along with Lewi Tonks, also discovered that an inert gas improved the lifetime of incandescent globes. He also found that twisting a tungsten filament enhances efficiency.

Fritz Pfleumer 1881-1945
magnetic tape for sound recording

He invented magnetic tape for sound recording in 1927 and received a patent for it in 1928. He used paper and iron oxide, with lacquer as an adhesive to bind the oxide to the paper. In 1932, he granted rights to this invention to AEG. They used it with the first practical tape recorder, the Magnetophon K1, demonstrated in 1935.

Cpt Henry Joseph Round 1881-1966
LEDs, vacuum tubes

He contributed to vacuum tube development and developed a triode around the same time as Lee de Forest. He discovered feedback in vacuum tubes independently of Alexander Meissner and Edwin Armstrong.

He made the first report of what we now know to be a light emitting diode (LED), utilising 'cat's whisker' detectors, the first type of semiconductor detector, made of silicon carbide and producing faint yellow light.

Next month

That's all we have room for in this issue. As mentioned earlier, we will pick up where we left off in the fourth article next month, finishing our chronological list of inventors. We will then turn to discoveries credited to groups of people.

PE



Fig.41: the Bélinographe used a photocell to scan and transfer photos in 1907. Source: <https://w.wiki/7DAk>



Fig.42: the Marconi-Stille tape recorder. Source: Birmingham Museums Trust – <https://w.wiki/7Dup> (CC-BY-SA-4.0).

Mains Power-Up Sequencer, part 2

PART 2 OF JOHN CLARKE'S

MAINS POWER-UP SEQUENCER



This Sequencer solves problems that can occur when switching on multiple mains-powered devices, like circuit breakers tripping or loud thumps from speakers. It can also be used as a master/slave power-saving solution. The Sequencer can handle up to four devices but multiple units can be chained to handle 8, 12 or more.

The Sequencer can switch on one to four (or more) devices in sequence, with an adjustable delay between each power-on. It can also switch them off in sequence, either in the same order as they were switched on or in the reverse order.

It can be configured to start to switch on the devices in one of three ways: immediately when power is applied to it, when the appliance plugged into the first outlet starts to draw power (in which case the first outlet is always on), or when a separate, isolated mains supply comes online.

That last feature can join multiple Sequencers to control more than four devices. It can even allow you to switch on devices in sequence across multiple mains phases (eg, if you have a big lab full of equipment).

In last month's first article, we described all its features and how the circuit works. Now we pick up where we left off and move on to building it, followed by testing and configuration.

Construction

Most of the Mains Power-Up Sequencer's parts are assembled onto a double-sided PCB coded 10108231 that measures 203 × 134mm. The completed assembly is housed in an ABS or polycarbonate plastic IP65 sealed enclosure measuring 222 × 146 × 55mm.

Figs.5 & 6 show where all the components go on the circuit board.

You will not fill the entire PCB with components when building the Mains Power-Up Sequencer.

Typically, you would only install the Current Detection section or the Mains Input Detection section, but not both. Or you could decide not to use either, in which case none of those parts are needed. The parts list last month separated out the parts for the optional sections.

The OUT1 channel must always be installed, but note that there are a couple of component value changes in that section depending on whether Current Detection is installed.

Additionally, if Current Detection is not used, the two pads for CON7 must be connected using a short length of 10A mains-rated wire.

Before construction, you will need to decide on how many outlets you will install. The PCB is initially set for four outlets with the RA0 and RA1 pins

on IC9 tied to the 0V supply by short tracks on the underside of the PCB.

To change this, the bottom layer tracks right next to the RA0 and/or RA1 pads will need to be cut (eg, using a sharp hobby knife) and then those pad(s) soldered to the small adjacent pads on the top layer that connect to +5.1V.

Refer to Table 1 to see which need to be changed for one, two or three outlets. If you can't get the solder to reach across the gap, use a short length of component lead offcut.

Ensure you've properly isolated the pads before soldering them to those top pads, or you could short out the 5.1V supply (which will prevent the unit from working but shouldn't blow anything up).

Circuit sections

The Mains Power-Up Sequencer PCB screen printing separates the four mains output circuitry sections (OUT1 to OUT4) using lines to delineate each channel. The Current Detection and Mains Input Detection sections are also marked on the screen printing and in Fig.5, so it is easy to see where the components associated with each section are located.

Before construction, decide which sections you need using the information above. You can then start by installing the smaller ¼W resistors.

Warning: Mains Voltage

All circuitry within the Mains Sequencer operates at Line (mains) voltages. It would be an electrocution hazard if built incorrectly or used with the lid open. Only build this if you are fully experienced in building mains projects.

Constructional Project

They have colour-coded bands indicating the values (shown in the parts list last month), but it's best to use a digital multimeter (DMM) to check each resistor before soldering it in place.

Zener diodes ZD1 and ZD2 (if used) and TVS1 (if used) can also be installed now, taking care to orientate the zener diodes correctly. TVS1 can go in either way around.

Mount the ICs now, including the opto-couplers, taking care to get the correct IC in each place and with the proper orientation. We used sockets for IC9 and IC10, although you could solder them directly to the PCB, assuming that IC9 has already been programmed. The opto-couplers (IC1-IC8 and IC11) are not all the same, so don't get them mixed up.

Note that on the PCB, pin 5 of the IL410/4108 and the IL420/4208 have only a tiny pad for an increased separation distance between the internal Triac pins located at pins 4 and 6. Those pins are not connected to the rest of the circuit but you can solder them if you want to.

The Triacs can be mounted now. There are a few different ways to do this. One is to smear a thin layer of flux paste onto the large pad, then position the device on the PCB and solder one of the small leads. Check its alignment and, if it's OK, solder the other one. Otherwise, reheat the initial joint and nudge it into position first.

Finally, turn up your iron and feed solder slowly into the large tab, as

Table 1 – number of outlets

RA1 (pin 18)	RA0 (pin 19)	# outlets
0V (bot)	GND (bot)	4 (default)
0V (bot)	5.1V (top)	3
5.1V (top)	GND (bot)	2
5.1V (top)	5.1V (top)	1

it will take a while to melt. Once it gets hot enough, solder all along the exposed portion of the tab. The flux paste underneath will pull solder under the tab and solder it to the circuit board.

Alternatively, it is possible to tin both the pad and the tab of the device, clamp them together while heating the tab and feeding in more solder to reflow them together, then solder the two smaller pins.

Bridge rectifiers BR1 (and BR2 if used) can now be mounted. These components must be correctly orientated with the + lead inserted into the position marked with a + and seated close to the PCB before soldering.

The 1W resistors can be fitted now. Ensure the correct values are used and note that for the OUT1 channel, R1 is 470Ω 1W when the Current Detection components are installed or 330Ω 1W when the Current Detection circuitry is not installed.

There are 1MΩ resistors under the relays that are inserted from the underside of the PCB, as shown in Fig.6. Solder these in place

and cut the leads flush with the top of the PCB. Then mount the 1kΩ 5W resistor with a gap of about 1mm from the PCB, to allow air to circulate.

Next, fit the capacitors, of which there are three types: the mains X2-rated capacitors, electrolytic capacitors, and MKT polyester types. The electrolytic capacitors need to be orientated correctly since they are polarised, while the others can be installed either way around.

For the OUT1 channel, C1 is 220nF X2 when the Current Detection components are installed or 10nF X2 when they are not installed. We have provided for the different sizes and lead spacing on the PCB.

Next, install potentiometer VR1 and the three toggle switches S1-S3. Then, mount the current transformer, T1, if used.

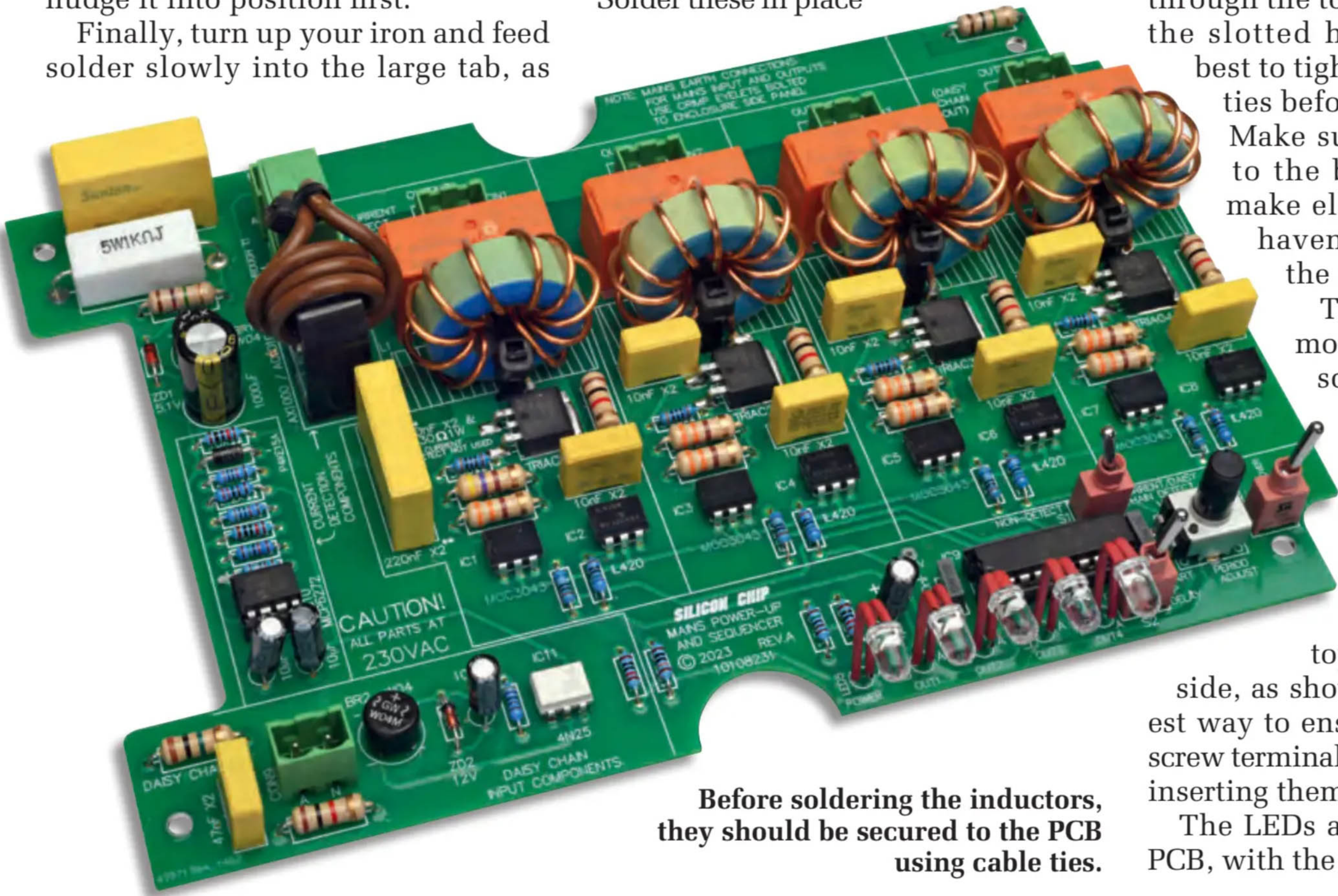
Winding inductors L1-L4

It's much easier to mount inductors L1 to L4 before the relays. These are wound using a 500mm length of 1.25mm diameter enamelled copper wire, with 10 turns evenly spread around the powdered iron toroid. Strip the insulation back by 1mm at each end of the wire using a sharp craft knife, insert the wire ends into the holes allocated and solder them in place from the top side of the PCB.

Each inductor is supported using a 200mm-long cable tie that loops through the toroid and then through the slotted holes in the PCB. It's best to tighten and trim the cable ties before soldering the leads. Make sure the solder adheres to the bare copper; it won't make electrical contact if you haven't fully stripped back the enamel.

The relays can now be mounted, followed by the sockets for the two-way terminal blocks. They must be inserted so the plug-in screw connectors are orientated correctly, with the screw head access positioned toward the top edge of the PCB (left side, as shown in Fig.5). The easiest way to ensure this is to plug the screw terminals into the sockets before inserting them into the PCB.

The LEDs are mounted above the PCB, with the leads bent by 90° 4mm



Before soldering the inductors, they should be secured to the PCB using cable ties.

Mains Power-Up Sequencer, part 2

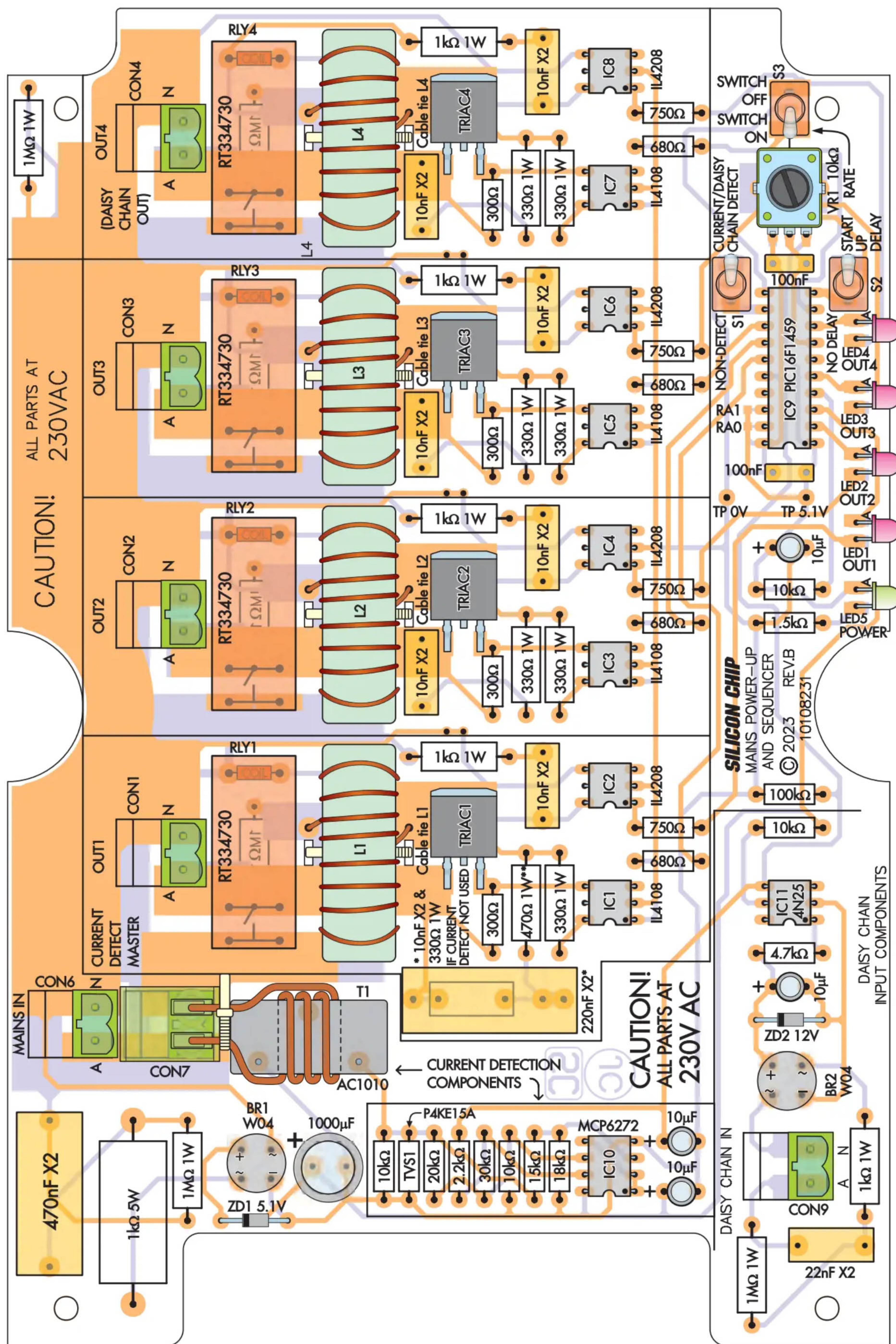


Fig.5: the PCB is divided into sections by lines. All components outside the boxed sections should be fitted, along with the OUT1 section and however many other outputs you need. Depending on how you plan to use it, you can also add either the Mains Input Detection ('daisy-chain') components or the Current Detection components (including T1), but not both.

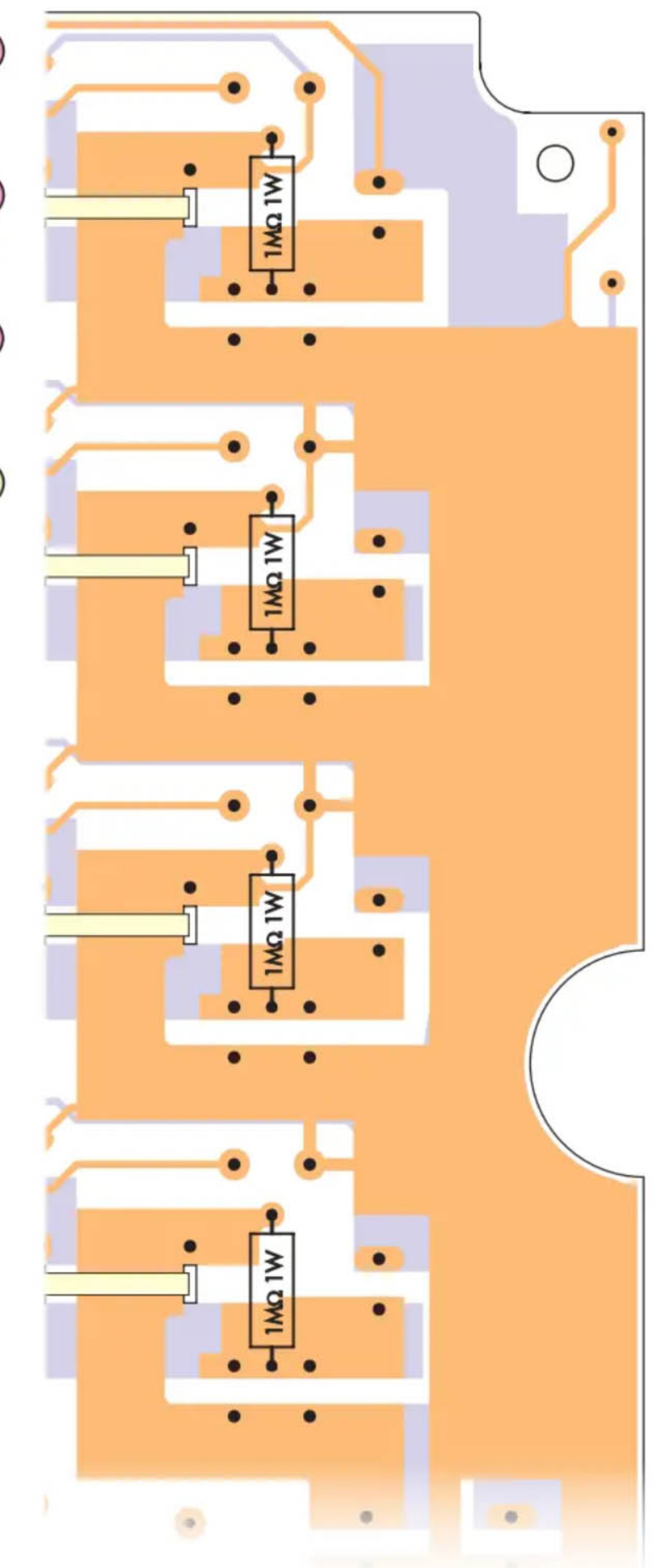


Fig.6: the only components you need to fit on the underside of the PCB are these four 1MΩ resistors underneath the relays. You can omit those from any output sections that are not being populated. This diagram is shown at 70% of actual size.

from the rear of the LED, so they sit horizontally. First, cover each lead with a 20mm length of 1mm diameter heat-shrink tubing. Then shrink the tubing with a hot air gun and bend the leads, ensuring that the anode (longer lead) will be orientated correctly with the LEDs bent (anodes facing to the top in Fig.5).

The LEDs stand 20mm above the

PCB when measured from the top surface of the PCB to the LED centreline.

Case preparation

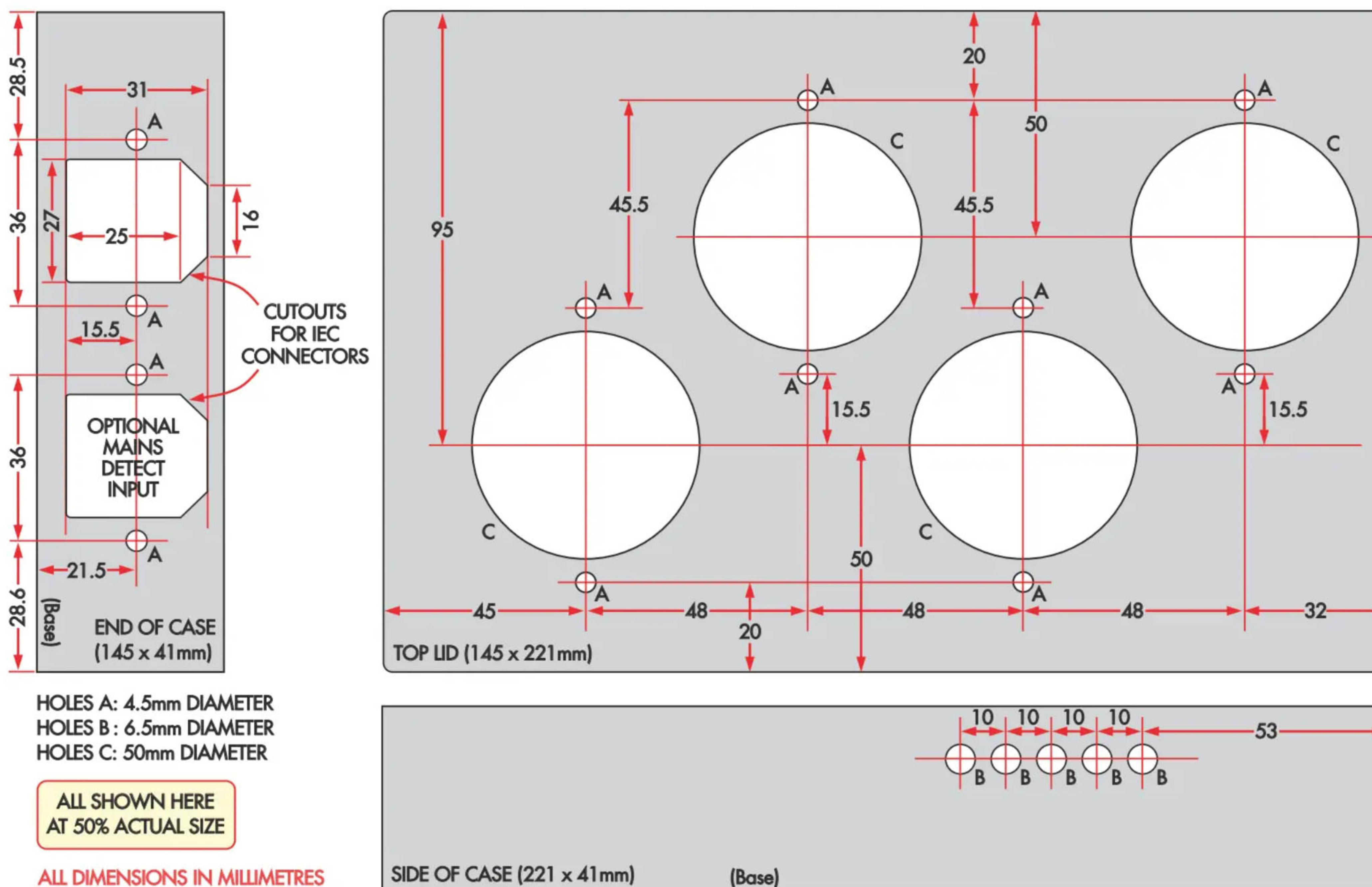
Before attaching the PCB, the IEC connector cutouts must be made in the side of the enclosure. You will also need to drill holes in the lid for the GPO sockets and in the enclosure side for the LED indicators. The re-

quired holes are shown in Fig.7. It can be downloaded as a PDF from our website at 100% scale and printed at actual size to use as a template.

Don't make the holes in the lid just yet as there are some options there, which we'll get to shortly.

Additionally, the two plastic stand-offs (not the ones with brass tapped inserts) that would be beneath the

Constructional Project



OUT4 components on the PCB need to be shortened using a large drill to allow clearance for soldered joints under the PCB.

Wiring

You can install the mains outlets in one of two ways. One way (as in our prototype) is to use surface-mounting GPOs on the lid of the enclosure, as shown in Fig.8.

Alternatively, you can use inline mains sockets and mains leads (possibly cut from extension cords), held

in place using cord grip grommets on the side of the enclosure, as shown in Fig.9. In this case, the Earth wires are attached to an M4 bolt on the side of the enclosure.

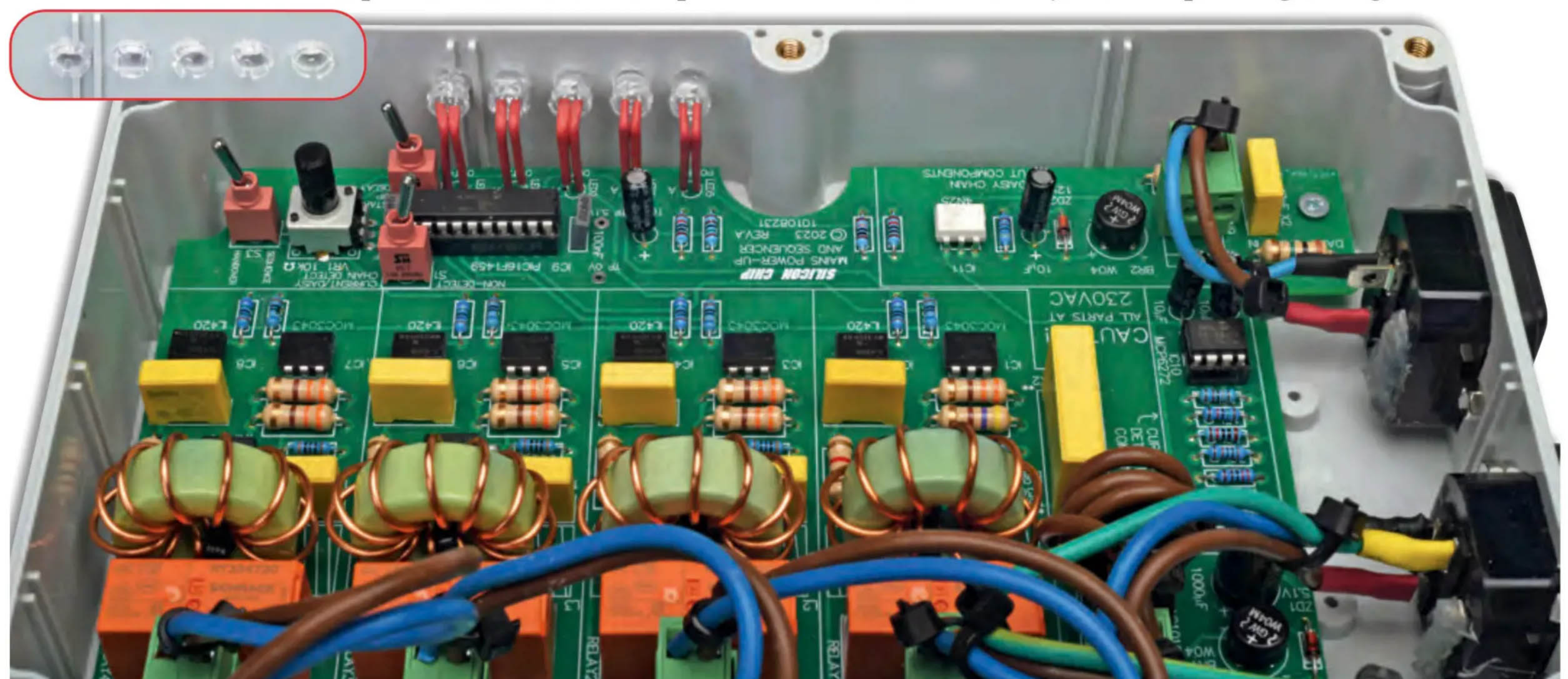
We provide cutout positions for the GPO sockets in Fig.7 since they need to be positioned on the lid so they don't foul PCB components underneath. We haven't provided drilling details for the alternative method using the cord grip grommets as the positioning is not so critical.

However, the cutout shape for

cordgrip grommets is important as it needs to be made so the grommet fits snugly when the cord is captured, so the lead cannot be pulled out from the grommet. The cutout shape is essentially an elongated circular hole.

Cable glands could be used instead of cordgrip grommets. In that case, it is essential to secure the gland nut so that the mains cable cannot be pulled out. This can be done by coating the gland threads with superglue before tightening the nut to secure the mains cable lead.

The LEDs are inserted into 16kV-rated bezels mounted on the side of the case to prevent shock hazards; how to mount the bezels is shown in the inset photo. The switches and potentiometers are used to adjust the sequencing settings.



Mains Power-Up Sequencer, part 2

The large cutouts for the mains GPO sockets and IEC connectors can be made by drilling a series of small holes around the inside perimeter, knocking out the centre piece and filing the outline to a smooth finish. Other methods include using a hole saw or speed bore drill to remove most of the inner area and then filing the rest to the shape required.

Once the drilling and filing are complete, install the IEC connector(s). The PCB can then be placed inside the case, and the LEDs inserted into the bezels as you drop the PCB into the enclosure. Then secure the PCB to the base of the enclosure with 6mm-long M3 machine screws into the case's integral brass inserts.

We specify Cliplite bezels specifically since they cover the LEDs and are rated to withstand 16kV, so they protect against a possible shock hazard should the LEDs fail. Using exposed LEDs at mains potential could be an electric shock hazard. Most 5mm LEDs don't specify the insulation capability of the package between the LED dome and the LED die inside. So use the bezels specified to ensure safety.

The IEC connector must be secured using countersunk 10mm Nylon M3 screws, although you can use metal nuts. The Nylon screws are essential as they avoid the possibility of the screws becoming live (at mains voltage) should a mains wire inside the enclosure come adrift and contact a screw holding the IEC connector.

Before attaching the mains GPO outlets and LED indicators, you can download and print out the front panel label shown in Fig.10. Details on making a front panel label are at siliconchip.au/Help/FrontPanels

The download includes two versions of the front panel. One front panel version does not have labelling for the Mains Detect Input IEC connector if you haven't installed it.

All wiring must be run as shown in either Fig.8 or Fig.9, using mains-rated cable. Be sure to use 10A wire (7.5A is OK for the Mains Detect Input wiring). The brown wire must be used for the Active wiring, blue for Neutral and green/yellow striped for the Earth wiring.

Note again that if you are not installing the Current Detection, then the two pads for CON7 need to be joined using 10A mains wire (ideally brown).

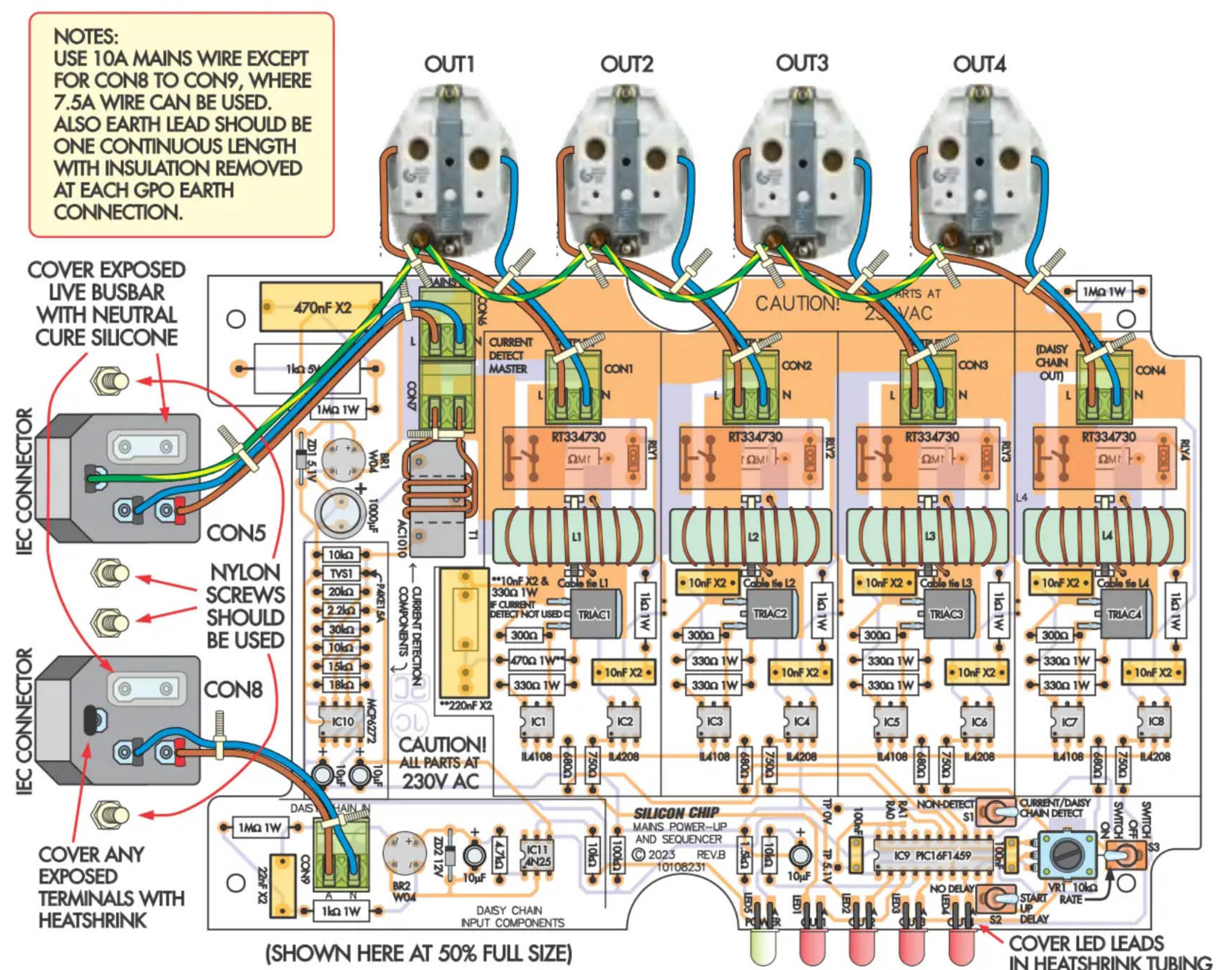


Fig.8: the wiring for the GPO version, which is what we built. Use 10A mains-rated wire with the correct colours for all connections, although the optional Mains Detect Input wiring can use 10A or 7.5A mains-rated wire. Don't skip the cable ties as they have an important safety function.

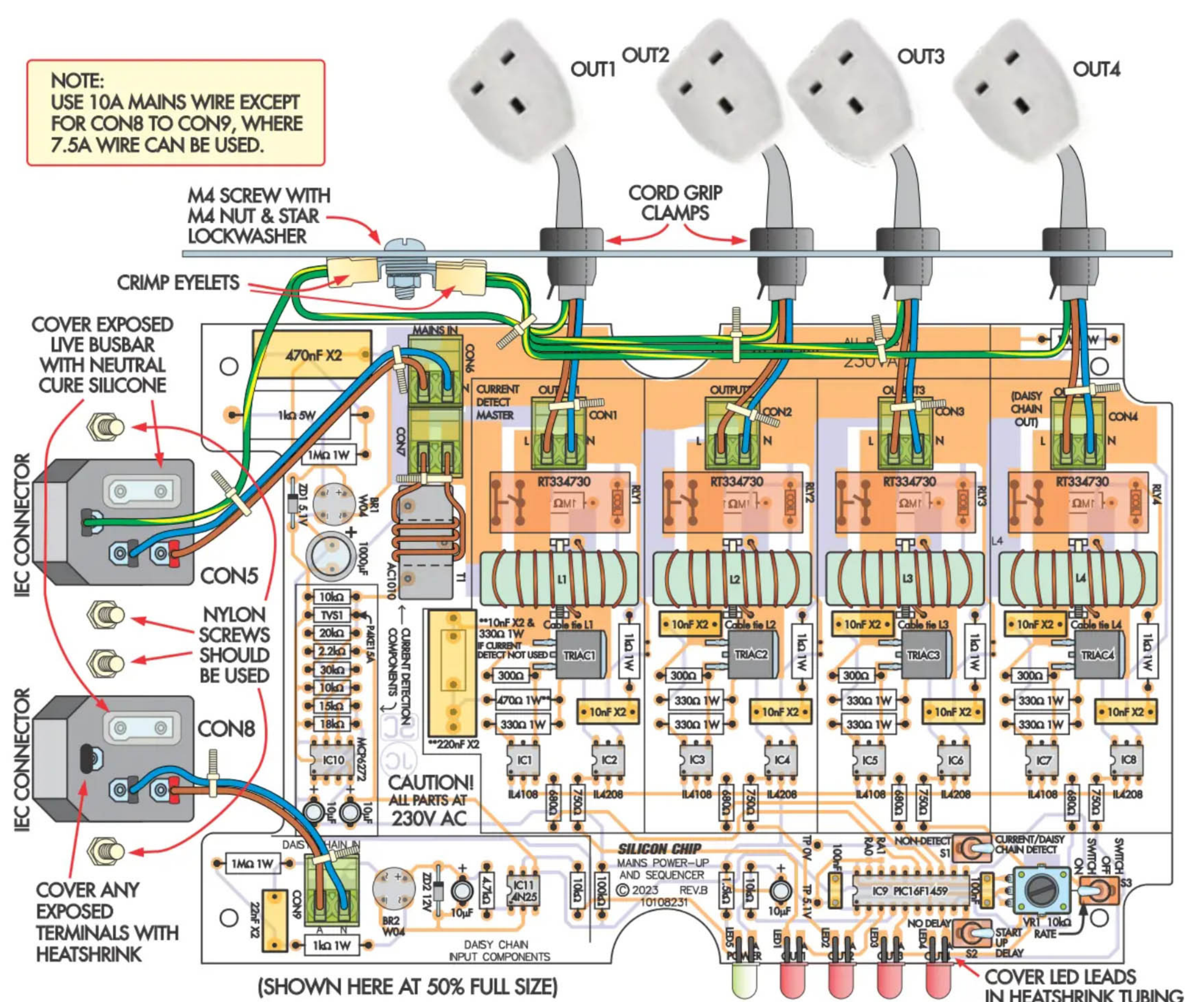


Fig.9: the wiring for the non-GPO version is similar to that shown in Fig.8 but the Earth wires are terminated slightly differently. The output cables can be made either by connecting mains flex to individual line sockets, or by cutting the plug ends off 10A extension cords.

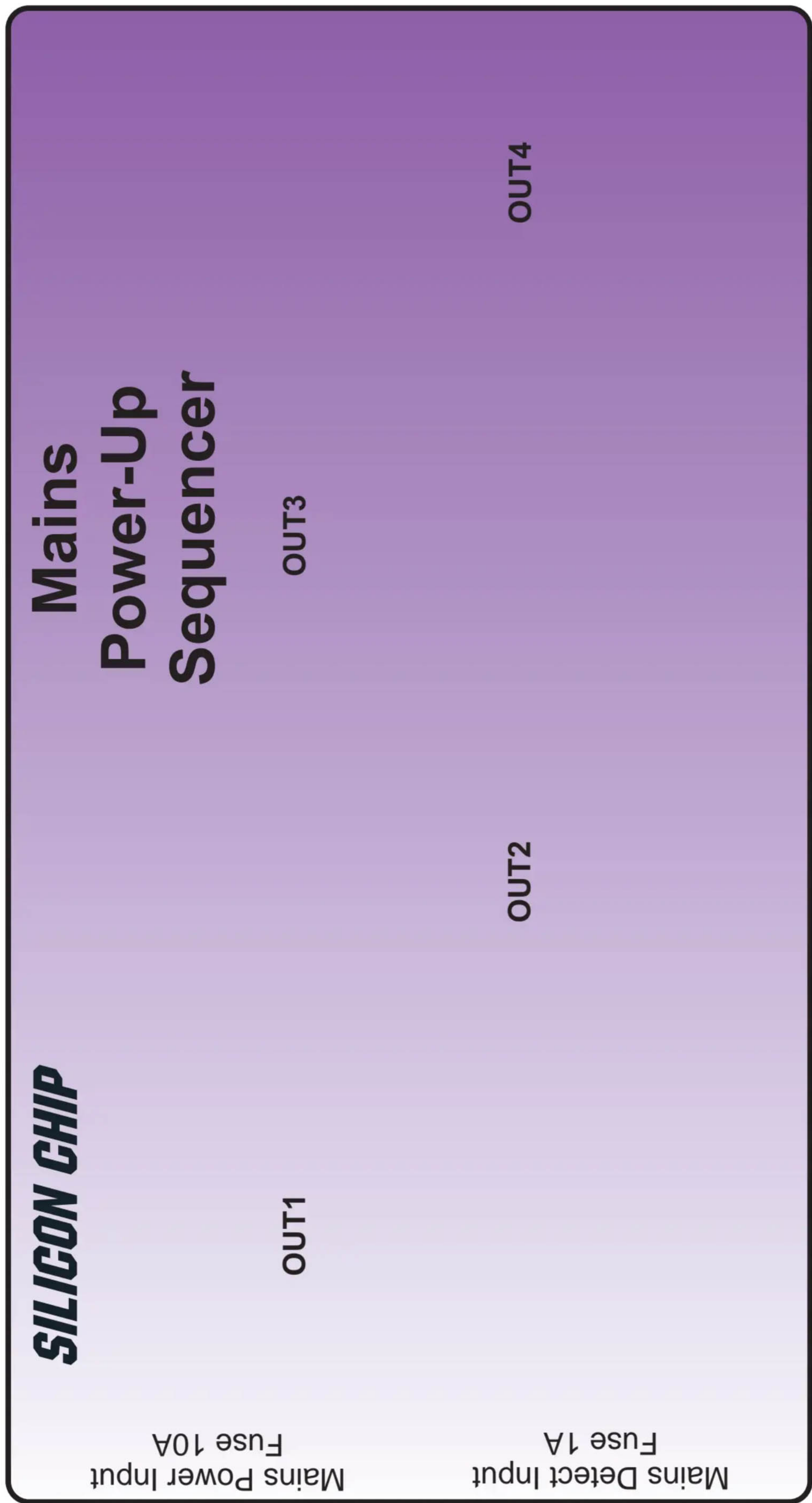


Fig.10: the lid label indicates which inputs and outputs have which function, while the side label shows what each LED means. There's another version of the label that you can download from our website without the text for the Mains Detect Input if you aren't using that feature.



For the lid-mounted GPOs, the Earth wire from the IEC socket must go straight to the first GPO Earth terminal, then to the second and so on as a single length of wire. To do that, strip the insulation off a single piece of wire at each connection point.

Take great care when making the connections to the mains sockets (GPOs). In particular, be sure to run the leads to their correct terminals. The GPO sockets will have the A, N and E clearly labelled, although Active might be marked with an L (Live) instead of an A. Do the screws up tightly so the leads are held securely. Similarly, ensure that the wires to the two-way screw terminals are firmly secured.

For the version without GPOs, the Earths are connected to crimp eyelets that are then all attached to the M4 Earth bolt, which is secured to the case using a star washer and nut.

Be sure to insulate all the Active and Neutral connections on the IEC connectors with heatshrink tubing for safety, and cable tie the wires as shown to prevent any broken wires from coming adrift. Use 5mm diameter heatshrink for the wires to the IEC connector.

Secure the Active and Neutral leads together using cable ties. Also, use neutral-cure silicone sealant (eg, roof & gutter silicone) to cover the Active bus piece that connects the Active pin to the fuse at the rear of the IEC connector. That bus is live, and there is no need to leave it exposed.

Testing

Always attach the lid using at least two screws at diagonally opposite locations before switching on the power. All the circuitry is operating at mains potential, so do not touch the components unless the power is off and the IEC power leads have been disconnected for at least ten seconds.

Before applying power, check your wiring carefully and ensure all mains connections are covered in heatshrink tubing and the wiring is cable tied. Then install the 10A fuse inside CON5's fuse holder and verify that IC9 is plugged into its socket and correctly orientated. If you have installed the Mains Input Detection circuitry, insert the 1A fuse into CON8.

VR1 can initially be set to mid-travel for a nominal 10-second sequence interval. If set fully anti-clockwise, VR1 gives a 100ms sequence delay period

Mains Power-Up Sequencer, part 2

while near full-clockwise (about 10° away) gives a 22 second sequence interval.

Set switch S1 to the left (open) position to disable Current Detection. Set S2 to the right for a startup delay and S3 to the left so VR1 sets the on-sequence period.

Remember the earlier advice to unplug the unit before opening the lid and adjusting any settings. Also note that settings like the periods are only stored at power-up. Making adjustments while the power is on won't do anything.

On power-up, check that the power LED lights and that the OUT1 LED lights after about ten seconds, followed by OUT2 after another ten seconds. The remaining LEDs should light after similar periods.

You can test the off-sequencing if you have installed the Current Detection or Mains Input Detection circuitry. To do this, unplug the unit, open the lid and move S1 to the right (closed) position. Reinstall the lid and power it back on.

If using Mains Input Detection, plug CON8 into the mains and the startup sequence should begin. Disconnect or switch off that supply and the LEDs should switch off in sequence, starting with the last output and finishing with OUT1. The default delay for the off-sequence is two seconds.

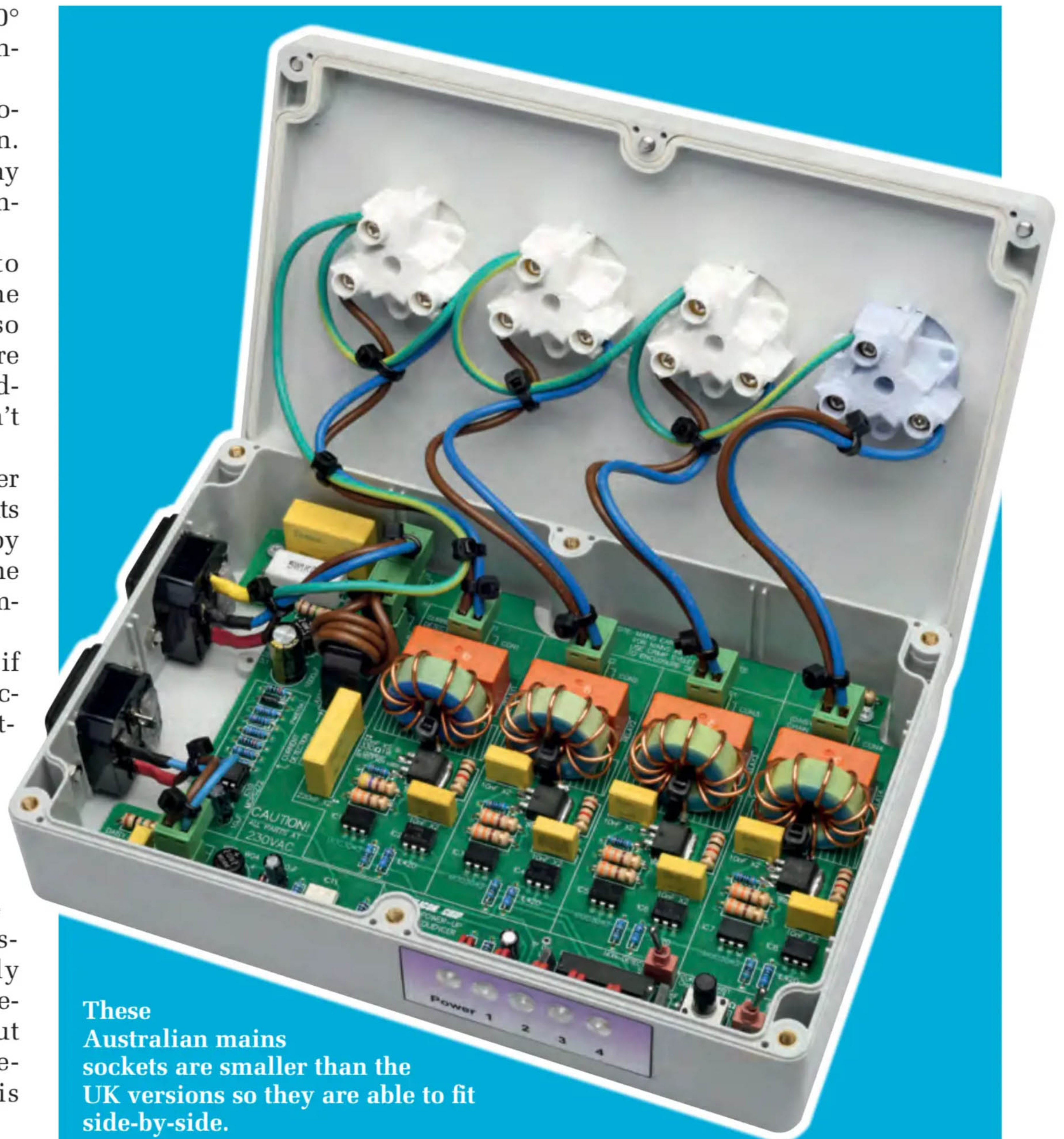
Alternatively, if using the Current Detection circuitry instead, plug an appliance into OUT1 and switch it on to trigger the on-sequence, then unplug it or switch it off to trigger the off-sequence. Again, the off-sequence should start with the last output and finish with OUT1.

Settings

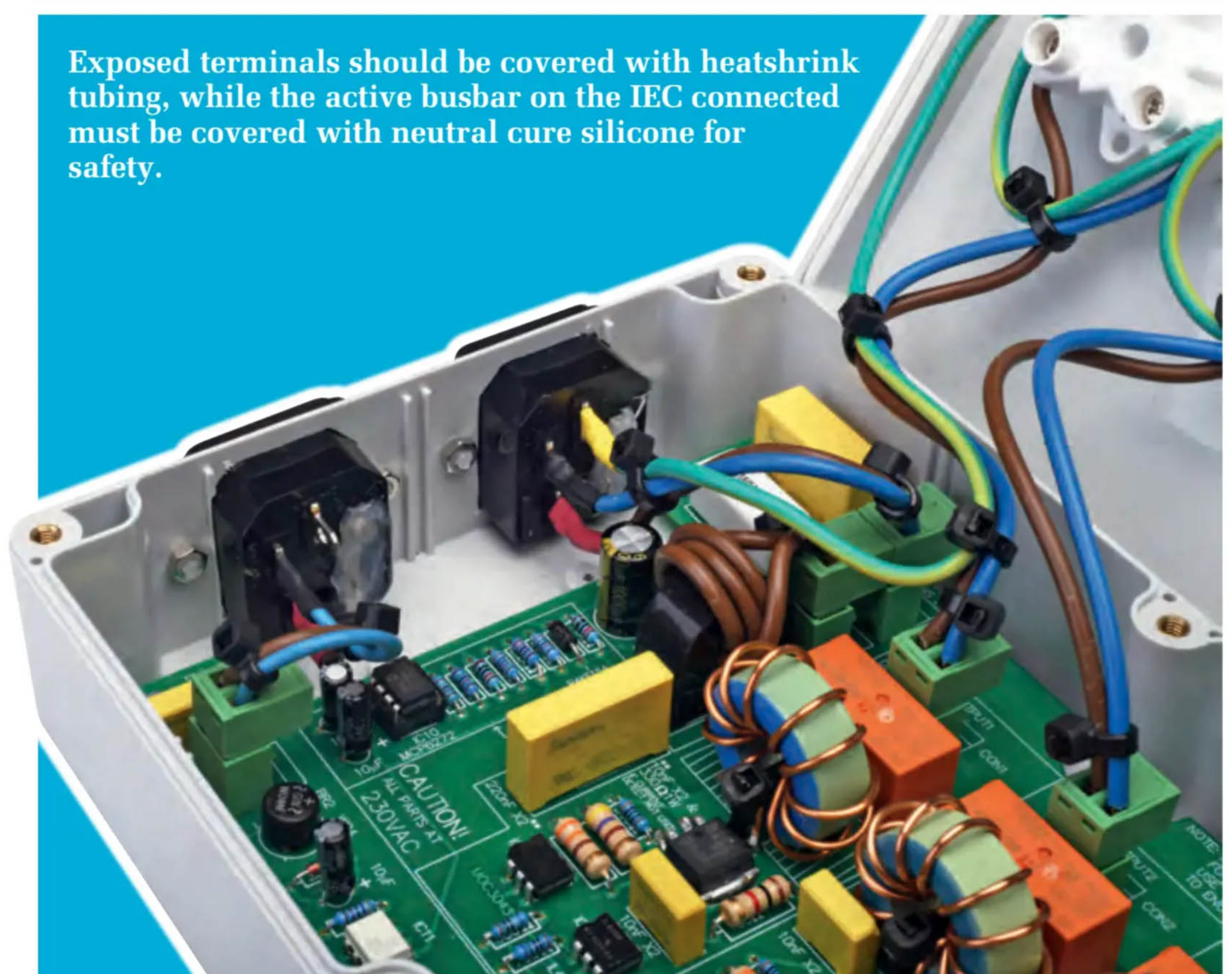
Two lots of settings can be made. First, there are the on-sequence and off-sequence periods, set using VR1. The on-sequence period is set with switch S3 in the left position and is only stored at the instant that power is switched on. To set the off-sequence rate, you also use VR1, but place S3 in the right-hand position before powering it up.

Each value is stored in flash memory, so it is recalled at power up, allowing you to set these two periods independently.

For these settings, VR1 can be adjusted from fully anti-clockwise to

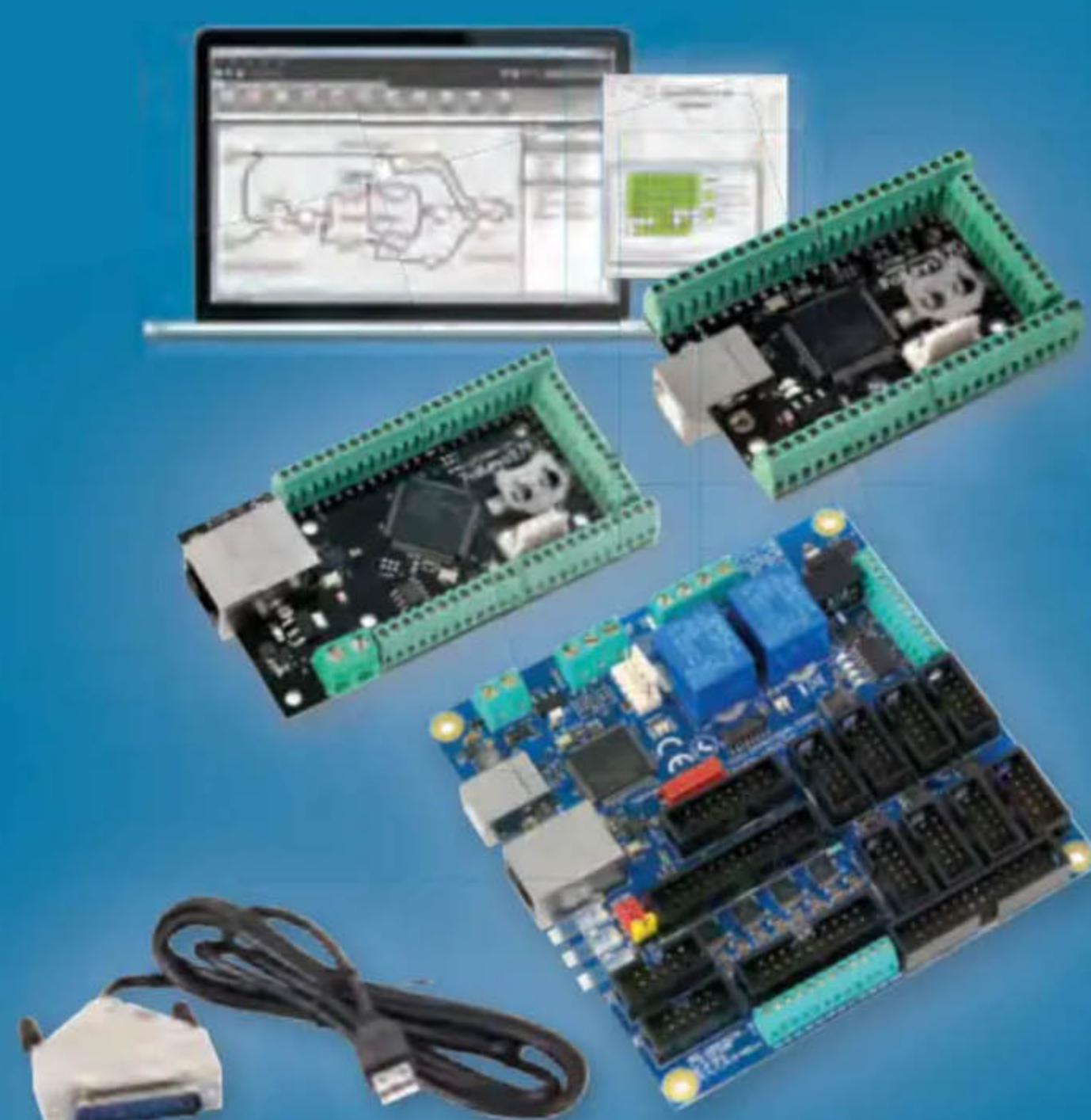


These Australian mains sockets are smaller than the UK versions so they are able to fit side-by-side.



Exposed terminals should be covered with heatshrink tubing, while the active busbar on the IEC connected must be covered with neutral cure silicone for safety.

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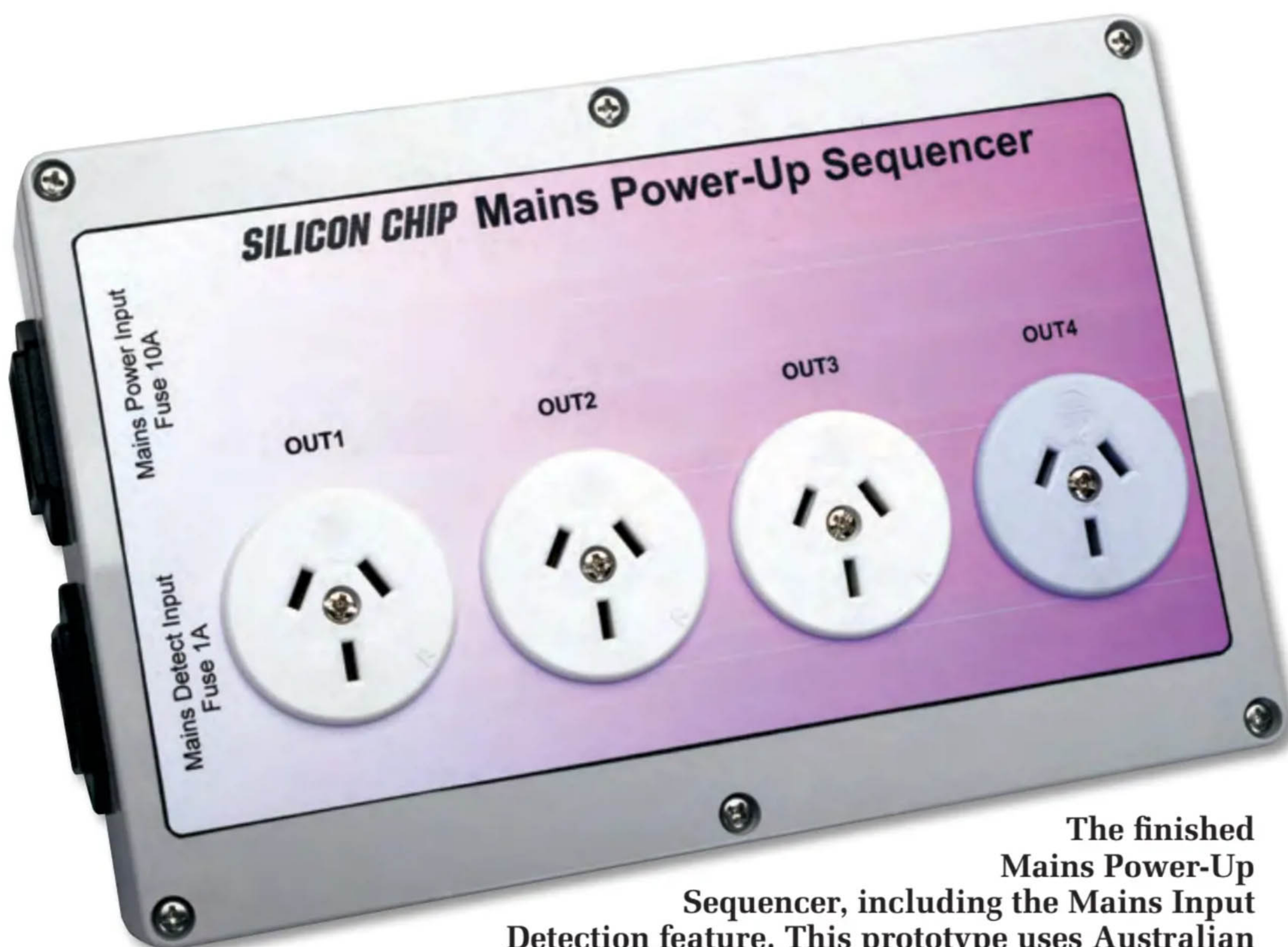
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Constructional Project



The finished Mains Power-Up Sequencer, including the Mains Input Detection feature. This prototype uses Australian sockets, but we have updated the drilling details and created a label to suit the larger UK BS1363 sockets, which won't fit side-by-side.

about 10° short of fully clockwise. That gives a range of 100ms (anticlockwise) to about 22s (near clockwise).

The other settings are made with VR1 set fully clockwise, which causes the Sequencer to enter another mode. It does two things in this position.

One is to measure the voltage from the precision rectifier when no appliance is connected to OUT1. This is the offset voltage from the op amp circuit, which is usually a few millivolts. This value is stored and subtracted from any future Current Detection measurements. If you are not using the Current Detection, it still happens but won't affect anything.

The other function of this mode is setting the off-sequence direction. With the power off and the unit unplugged from the wall, rotate VR1 fully clockwise. No appliance should be plugged into the sequencer GPO (OUT1) outlet or any mains power

applied to the Mains Detect Input (if used).

If switch S3 is set to the right, you will set the off-sequence to forward, meaning that OUT1 switches off first. If S3 is placed to the left, it sets the reverse off-sequence direction, so the last outlet switches off first. The initial setting of the programmed microcontroller is this reverse off-sequence.

After a few seconds in this mode, the Sequencer can be unplugged. After that, remove the lid and rotate VR1 back from fully clockwise to the desired period for the sequence rate, depending on the position of S3. This is important as, if VR1 is left set at the fully clockwise position, the Sequencer will not run to switch on any outlets.

Table 2 summarises the functions of switches S1, S2, S3 and potentiometer VR1. Settings are only changed at power-up. **PE**

Table 2 – power-up settings

Switch	Left (open)	Right (closed)	VR1
S1	No Mains/Current Detection	Mains/Current Detection enabled	
S2	No initial delay	Delay before on and off sequences	
S3	VR1 sets on-rate	VR1 sets off-rate	100ms to 22s (from full anti-clockwise to 10° less than clockwise)
S3	Reverse off-sequence	Forward off-sequence	Fully clockwise (also stores full wave rectifier offset)

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Harold S. Black, Negative Feedback and the History of Operational Amplifiers

Op amps and negative feedback circuits are ubiquitous today, and you would be forgiven for thinking that they have been around forever. But there was a time when electronics was still developing, and such devices had not yet been invented. That changed in 1927 with the bright idea of one clever man...

by Roderick Wall & Nicholas Vinen



One of the most significant early circuit ideas was Harold Steven Black's invention of negative feedback. In 1927, Harold S. Black (1898-1983) was on a ferry heading towards his office in the West Street Labs of Western Electric, the forerunner of Bell Telephone Laboratories in New York City.

An idea popped into his head that would dramatically change electronic communications, which continues to be used to the present day.

His idea was for a negative feedback amplifier, where the gain is accurately set and distortion limited by feeding part of the output signal back into the amplifier.

Black sketched his idea on a misprinted page of his copy of the New York Times, the only paper that he had on him. When Black got to his office, he had a colleague witness and sign it – see Fig.1.

Black's job had been trying to improve three- and four-channel telephone amplifiers based on carrier telephony for the last six years. For long-distance telephone calls, repeaters had to be added to cover the distance. But these repeaters had too much distortion, so by the time the audio signal reached its destination, it was unintelligible.

Black realised that amplifier distortion and noise could be reduced using negative feedback, at the expense of reduced amplifier gain. He later said that he did not know what made his idea pop into his head; it just came.

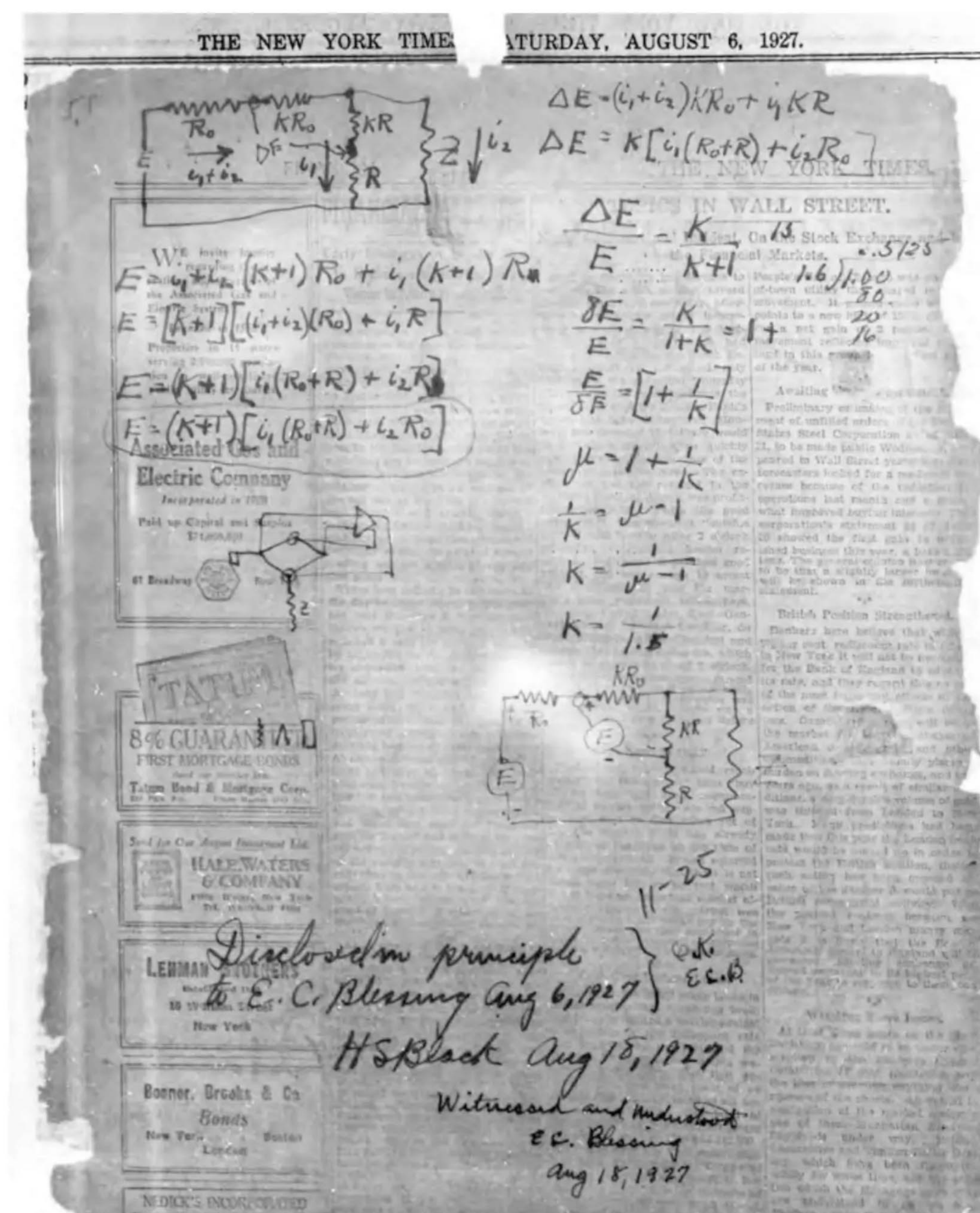
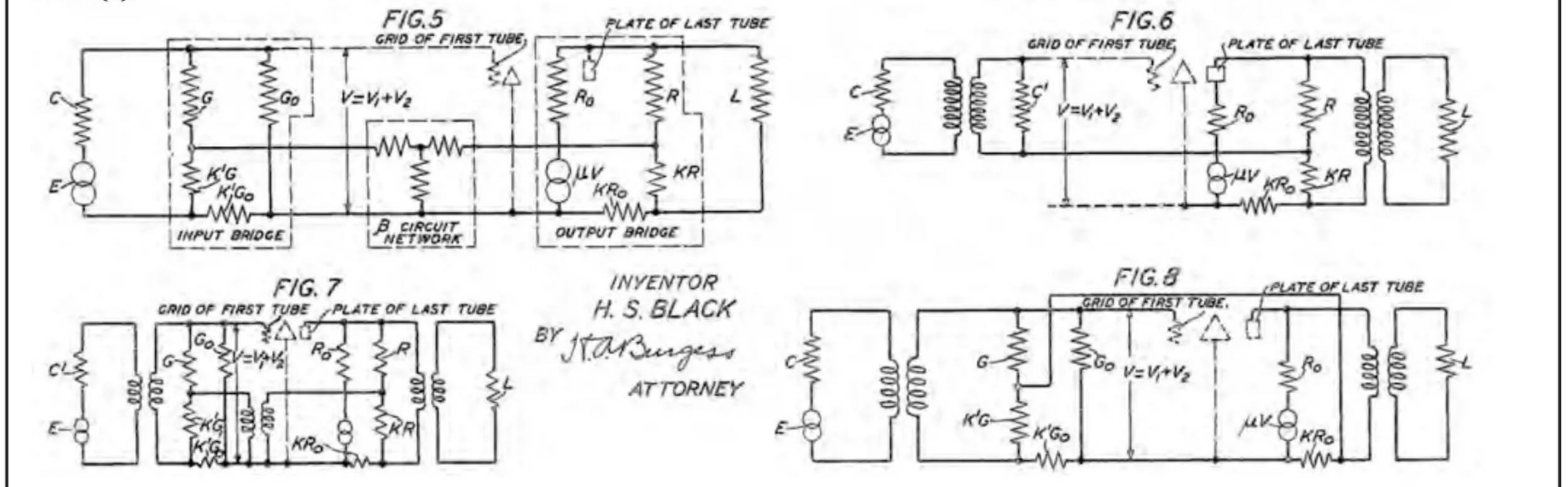


Fig.1: Harold Black's original hand-written notes on the principle of using negative feedback for distortion cancellation.

All About Op Amps

Fig.2: a page from one of Harold Black's many patents regarding negative feedback. This one is from patent 2,102,671, showing some possible ways of building an amplifier with negative feedback using valve(s).



Black used his new idea to design low-distortion broadband repeater amplifiers that were finally suitable for long-distance telephone calls. That allowed more channels over a pair of wires.

His patents

Harold S. Black was granted 62 patents during his career, 18 of which relate to negative feedback; these are listed in Table 1. His most famous patent is number 2,102,671, which you can view at <https://patents.google.com/patent/US2102671A>

If you replace the number in that link with the other patent numbers (plus the "US" prefix), you can view the relevant PDF.

This patent, titled "WAVE TRANSLATION SYSTEM", was filed in 1932 and granted in 1937. It comes to 87 pages and includes many detailed drawings (including circuits and plots) and plenty of explanatory text.

One of the most important sets of circuit diagrams (but far from the only one!) in this patent, appearing on page four, is reproduced in Fig.2. It shows four different ways of implementing his idea using 'tubes' or valves, the technology of the day.

Other important plots in the patent include gain curves, stability criteria, equivalent circuits and several practical implementations of the technique.

The importance of negative feedback

Almost all analog equipment manufactured today uses negative feedback. This includes circuits that handle audio signals, analog video, motor control,

battery monitoring, instrumentation and sometimes RF too.

The principle is used in TVs, radios, computers, medical equipment, control circuits, measuring instruments and mobile phones. You would find it very hard to find an electronic appliance that does not use negative feedback.

You will see negative feedback being used with operational amplifiers and in discrete circuits in most issues of *Practical Electronics*.

Operational amplifiers

This paved the way to the development of operational amplifiers (op amps); essentially, a monolithic implementation of a circuit which applies negative feedback.

Thousands of different types of op amps are available to suit just about any application; low-power types, high-speed types, high-gain types, precision

types, singles, doubles, quads etc.

The term "operational amplifier" goes back to about 1943, when this name was mentioned in a paper written by R. Ragazinni with the title "Analysis of Problems in Dynamics". The paper was the work of the US National Defence Research Council (1940), was published by the IRE in May 1947 and is considered a classic work in electronics literature.

George A. Philbrick Researches introduced the K2-W valve-based general-purpose op amp in 1952, more than a decade before the first transistorised version appeared (Figs.3 & 4).

The first solid-state transistor was successfully demonstrated on December 23, 1947, but it took a while before transistors were in widespread use. The first series of solid-state op amps were introduced by Burr-Brown Research Corporation and GA Philbrick Researches Inc in 1962.



Fig.3: a popular early valve-based op amp, the Philbrick Research K2-W.

Table.1: Harold S. Black's patents relating to negative feedback (patent numbers are hyperlinks)

When filed	Serial number	When issued	Patent number	Title	Pages
UNKNOWN	UNKNOWN	7 February 1928	CA277770A	Wave signalling system	NA
8 August 1928	298,155	Split into serial numbers 411223 & 411224 below			NA
3 December 1929	411,223	21 December 1937	2,102,670	Wave Translation System	21
3 December 1929	411,224	4 June 1935	2,003,282	Wave Translation System	12
26 March 1930	439,205	NA	Not granted	Wave Translation System	NA
3 April 1931	527,371	1 August 1933	1,920,238	Wave Translating System	17
22 April 1932	606,871	21 December 1937	2,102,671	Wave Translation System	87
30 September 1932	635,525	28 May 1935	2,002,499	Wave Translation System	10
29 December 1932	649,252	20 August 1935	2,011,566	Wave Translation System	7
29 March 1933	663,316	9 July 1935	2,007,172	Wave Translation System	6
29 March 1933	663,317	27 September 1938	2,131,365	Wave Translation System	12
25 September 1934	745,420	16 November 1937	2,098,950	Vacuum Tube Circuit	5
6 October 1934	747,117	17 March 1936	2,033,917	Electric Wave Amplifying System	5
5 December 1936	114,391	27 September 1938	2,131,366	Electric Wave Amplifying System	5
5 December 1936	114,390	6 August 1940	2,209,955	Wave Translation System	29
23 March 1937	132,559	18 April 1939	2,154,888	Wave Translation System	5
10 November 1937	173,749	3 December 1940	2,223,506	Wave Amplification	7
27 May 1938	210,333	17 June 1941	2,245,565	Wave Translating System	5
20 December 1938	246,791	7 October 1941	2,258,128	Wave Translating System	9
30 July 1940	348,433	26 May 1942	2,284,555	Signaling System	8
28 February 1942	432,860	20 July 1943	2,324,815	Stabilized Feedback System	7

The first solid-state monolithic op amp IC, designed by Bob Widlar and offered to the public in 1963, was the uA702 manufactured by Fairchild Semiconductors.

But it required strange supply voltages such as +12V and -6V and had a tendency to burn out. Still, it was the best in its day, and sold for about US\$300 (a fortune today!). It was used mainly by the US military due to its high cost.

Then the uA709 from Fairchild Semiconductor was released in 1965. It was introduced at about US\$70, and was the first to break the \$10 barrier, then not much later, the \$5 barrier.

By 1969, op amps were selling for around \$2 each. From then on, multiple manufactures produced op amps in

many varieties, up to the present day.

One particularly popular model was the uA741, which has been improved since it was first introduced in 1968. Some variants of it, such as the LM741, are still being produced today! Its equivalent circuit is shown in Fig.5. Modern op amps mostly use the same principles, but differ in some implementation details, such as the method of internal frequency compensation.

One big benefit of the op amp is its flexibility. It can perform a wide range of analog 'functions' with the addition of a few passive components. These functions include signal mixing, amplification, filtering (low-pass, high-pass, bandpass, notch etc), integration, differentiation, multiplication, simulated inductance and more.

You can think of op amps as the building blocks for most analog circuits.

Negative feedback

So how is negative feedback used to control an op amp to reduce the distortion and set a fixed gain?

The output voltage of an op amp is the non-inverting input voltage minus the inverting input voltage times a large factor (in some cases, over one million). If we say the gain is exactly one million, this means that:

- If the + input is 100µV and the - input is 99µV, the output will be +1V.
- If the + input is 100µV and the - input is 100µV, the output will be 0V.
- If the + input is 100µV and the - input is 101µV, the output will be -1V.

From this, you can see that if the difference between the input voltages is more than a few microvolts, the output voltage will be 'pegged' at one supply rail or the other. So unless we are using the op amp like a comparator (a possible op amp function), the inputs will almost always be at a very similar voltage. The negative feedback is typically configured to ensure that this is the case.

Let's say we feed 10% of the output voltage back to the inverting input and apply 1V to the non-inverting input.

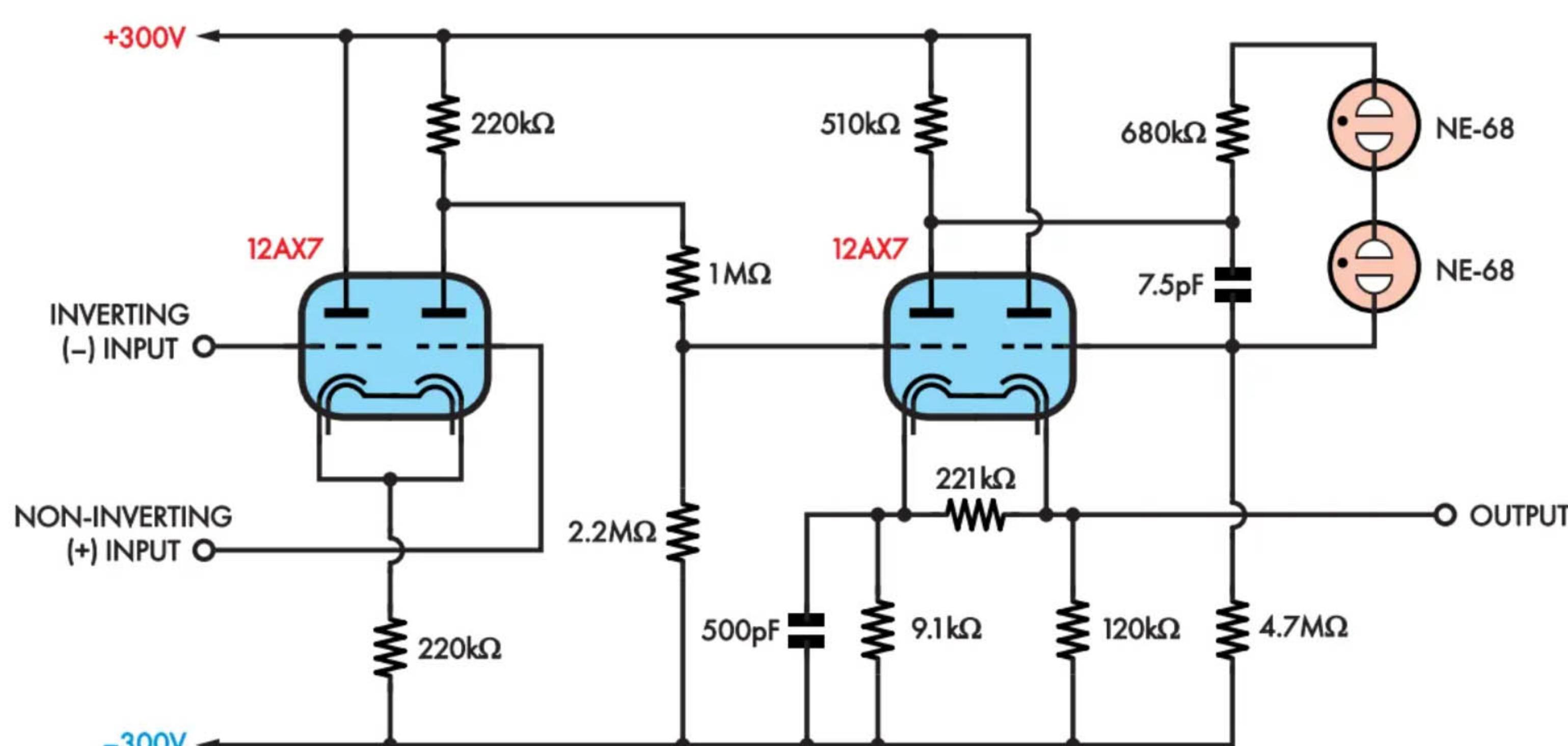


Fig.4: the K2-W uses a similar configuration to transistor-based op amps, with an input pair (one 12AX7 twin triode) followed by a voltage amplification/buffering stage made from another 12AX7 twin triode plus two neon lamps.

All About Op Amps

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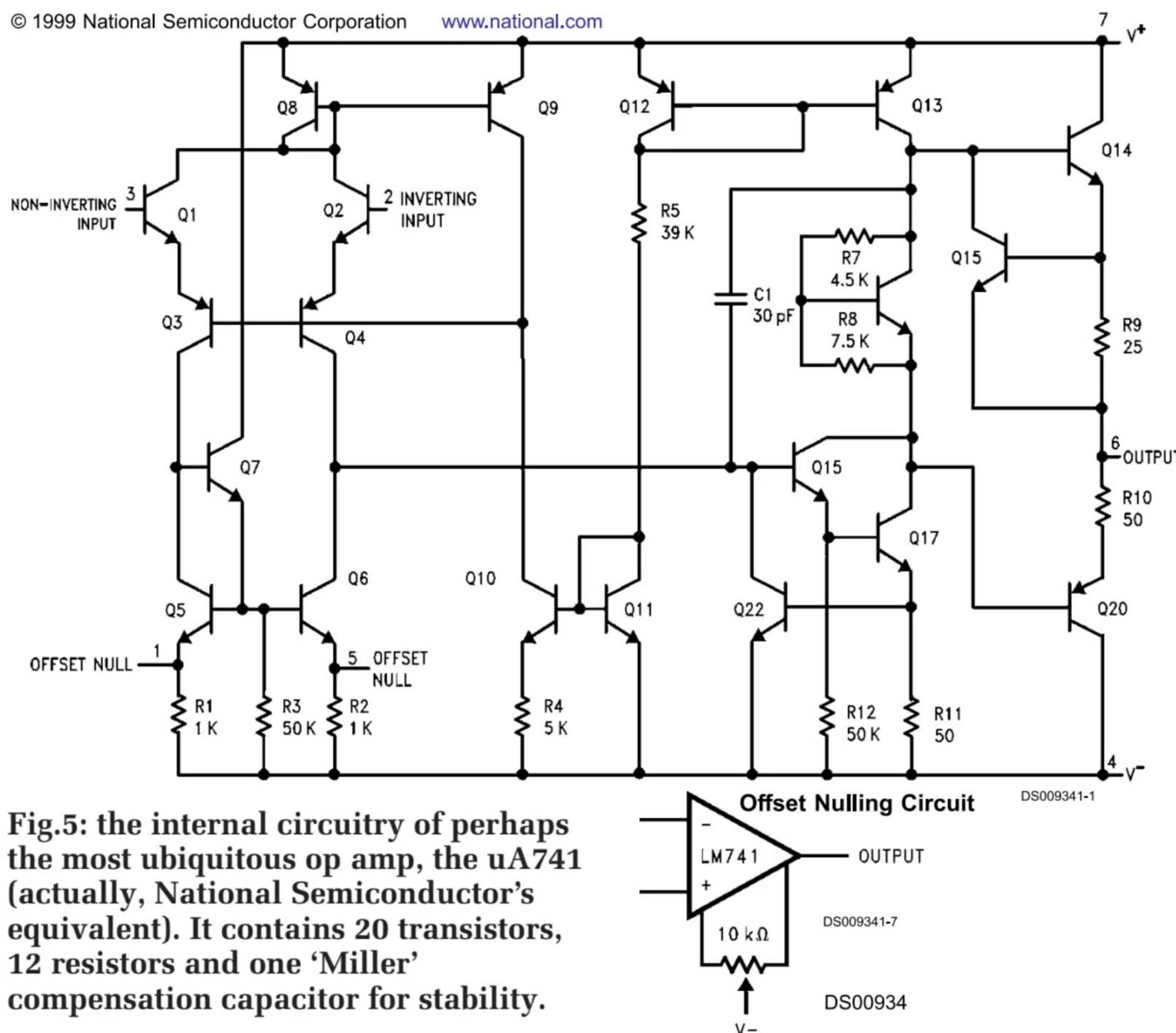


Fig.5: the internal circuitry of perhaps the most ubiquitous op amp, the uA741 (actually, National Semiconductor's equivalent). It contains 20 transistors, 12 resistors and one 'Miller' compensation capacitor for stability.

When the output is less than 10V, the voltage difference between the inputs will be positive, so the output voltage increases. When the output is greater than 10V, the voltage difference between the inputs will be negative, so the output voltage will decrease. Thus, the output voltage will tend towards 10V.

The only real sources of error in a DC context are the input offset voltage (the output not being exactly 0V with both inputs at the same voltage) and the finite gain, which adds a few additional microvolts of error. But that's just one part per million or so.

So it is pretty close to an ideal amplifier with fixed gain; that is certainly not the case with a typical single-transistor or single-valve amplifier! Due to manufacturing tolerances, it is challenging to set up (bias) a single transistor or valve to provide an exact gain. Even if you achieve it (eg, by trimming), it will likely change with temperature and over time.

Note how the exact gain of the op amp is not important; it only affects the (tiny) gain error. The overall gain is set by the feedback divider, usually made of resistors (and sometimes capacitors), so it's easy to set it close to the desired value. It can be trimmed to be almost exact if required, and it's unlikely to drift.

Negative feedback also gives close

to ideal results for AC signals, as long as they are well below the op amp's bandwidth (usually specified as gain bandwidth, which must be divided by the configured gain). Thus, an op amp-based amplifier can give an essentially flat gain curve across a range of frequencies, whereas a transistor or valve will typically be far from flat unless it is a special type.

Here are some basic op amp circuits:

1) Unity-gain buffer

Fig.6 shows an op amp arranged as a unity-gain buffer. The output is fed back to the inverting input, so the output voltage tracks the non-inverting input. As the output of an op amp has near-zero impedance (due to feedback), but the input has a relatively high impedance, this configuration is useful to avoid the circuit feeding the input from being loaded

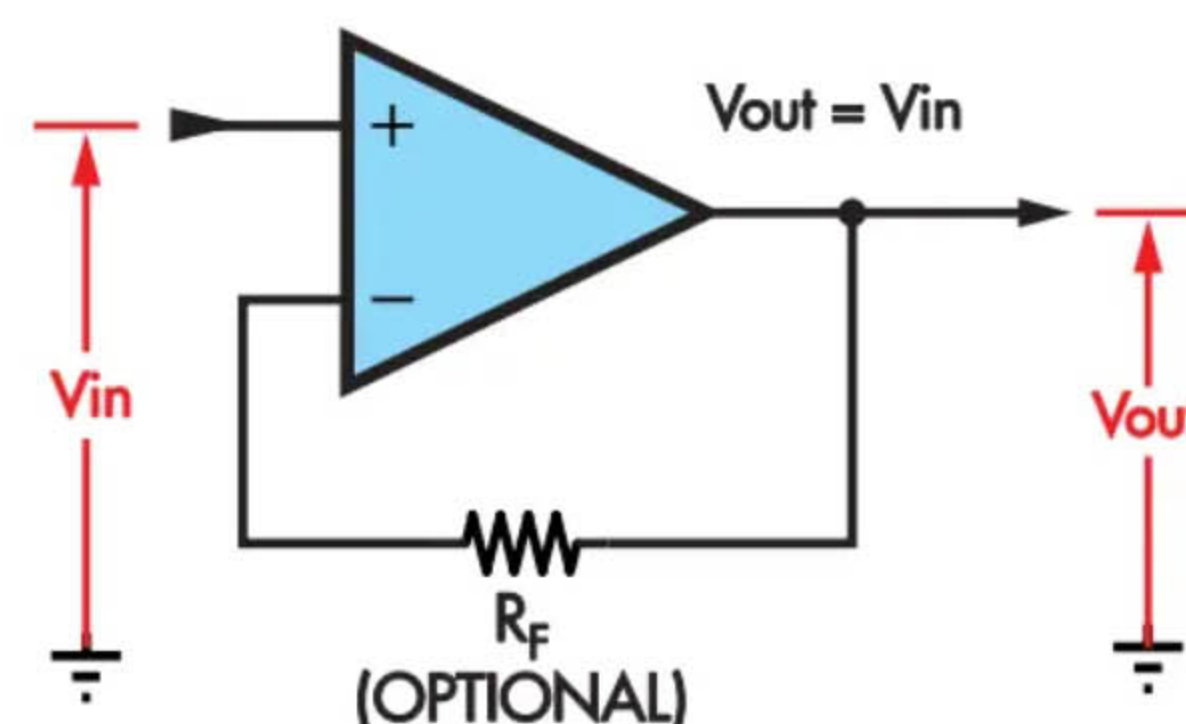


Fig.6: using an op amp to buffer a signal can be as simple as connecting its output to its inverting input. However, resistor R_f is a good idea to balance the input currents if the source impedance for the non-inverting input is relatively high.

by the circuit the output is driving.

Often, the output will be connected directly to the inverting input. But in some cases, the resulting source impedance mismatch between the inputs can cause temperature drift and other problems. Resistor R_f can be chosen to match the non-inverting source impedance to avoid this.

2) Non-inverting amplifier

Fig.7 shows an op amp providing non-inverting gain. The output voltage is an AC signal with the same shape as the input signal but an increased magnitude, by a factor of $R_f \div R_1 + 1$. As with the buffer, this circuit can be connected to a signal source that has a high impedance, but it still provides a low-impedance output.

Capacitor C_1 may be omitted, but it's usually a good idea to keep it. It reduces the circuit's gain at higher frequencies, thereby increasing stability and preventing the amplification of unwanted high-frequency signals.

You might see a high-value capacitor at the bottom of the feedback divider, between the bottom end of R_1 and ground, shown as an alternative connection for R_1 in Fig.7. This sets the circuit's DC gain to unity regardless of the AC gain, so it is mostly used when amplifying AC signals; also refer to Fig.19.

By reducing the DC gain of the circuit, it prevents the output from pegging at the positive rail on positive signal excursions, and also reduces the amplification of the input offset error voltage.

The practical gain limit depends on the op amp's gain bandwidth and the maximum signal frequency. For example, an op amp with a gain bandwidth

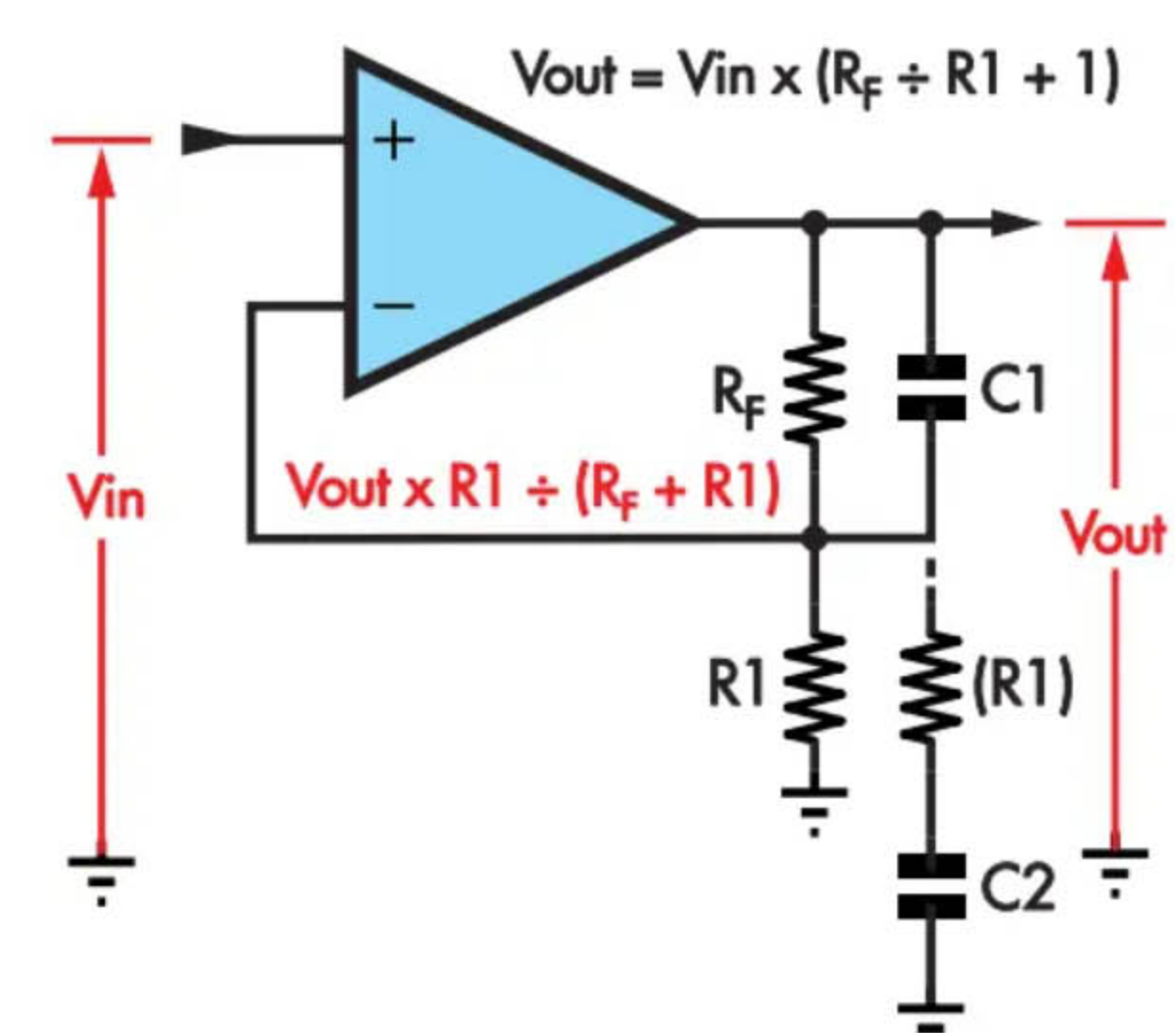


Fig.7: you only need two resistors to set up an op amp as a fixed gain voltage amplifier. As the signal is fed directly into the non-inverting input, the input impedance is high. Optional capacitor C_1 limits the bandwidth for stability, while C_2 can be used to reduce the DC gain to unity while having a higher AC gain.

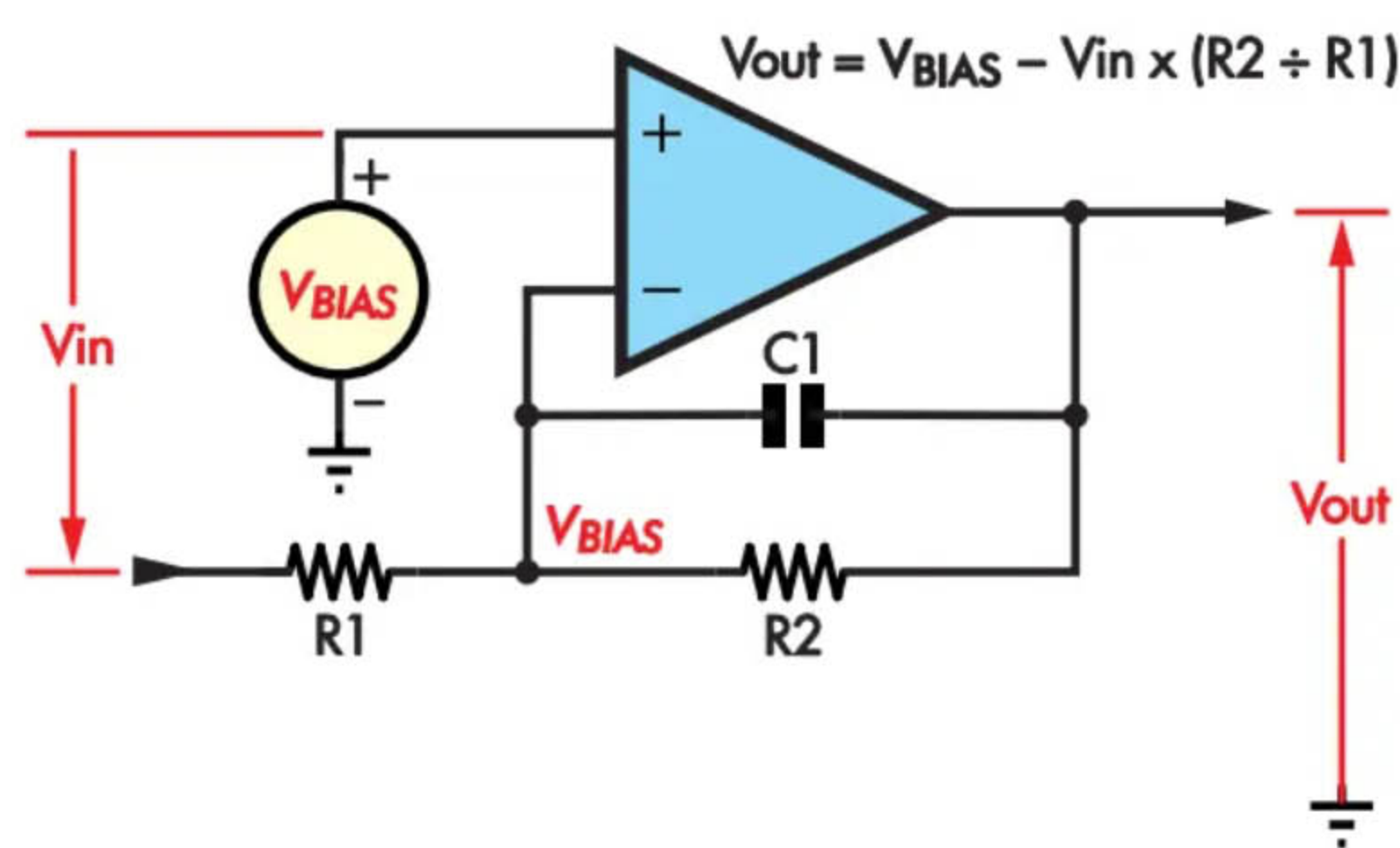


Fig.8: the inverting amplifier configuration also uses two resistors and one optional capacitor. While it has the advantage that the gain can be less than unity, the disadvantage is that the input impedance is equal to R_{IN} , rather than the usually much higher figure for the op amp's inputs.

of 3MHz has a maximum practical gain of 30 times for signals up to 100kHz (3MHz ÷ 100kHz). Noise and distortion in the output increase with gain, as there is less feedback (closed-loop bandwidth) for the op amp to work within.

3) Inverting amplifier

By feeding a signal into the inverting input rather than the non-inverting input, via a resistor, the signal is inverted and gain can still be applied, as shown in Fig.8. The gain is $-R_F \div R_{IN}$, so unlike the non-inverting version, gain values less than unity (ie, attenuation) are possible without a separate input attenuator.

An unfortunate consequence of this configuration is that the typically high input impedance of the op amp is reduced to the value of R_{IN} , so the circuit feeding the input is loaded more heavily. This can be solved by adding a unity-gain buffer between the signal source and the inverting amplifier.

One advantage of this configuration is that both op amp inputs are held at a constant voltage (V_{BIAS}), so there is no common-mode signal and therefore no common-mode distortion (often the dominant distortion mechanism).

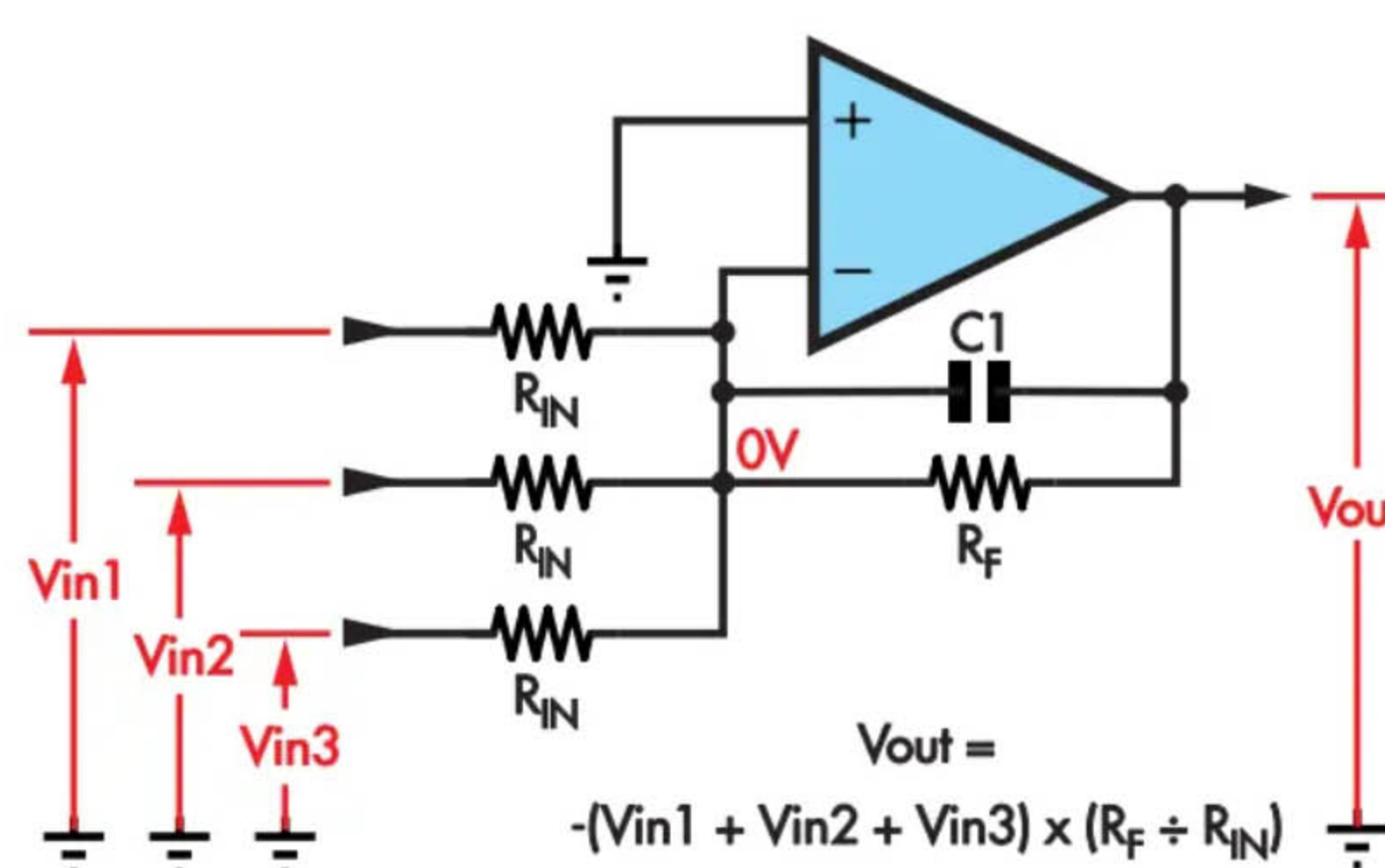


Fig.9: the virtual ground mixer is an inverting amplifier with multiple signal sources. As both op amp inputs are held very close to 0V, there is no way that the signals being fed in can interact with each other, except at the output of the mixer.

In this circuit, capacitor C1 performs a similar role as in Fig.7, although it is arguably more effective here since it reduces the gain at very high frequencies to zero rather than unity.

4) Virtual ground mixer

Fig.9 shows a circuit that is basically an inverting amplifier with multiple resistors feeding different signals into the inverting input. As the inverting input is held at a fixed DC voltage by the negative feedback, there is no possibility of cross-talk between the signals (which might be significant in a mixing console, where they are fed to multiple locations).

5) Differential amplifier

This is a very useful circuit used in many different forms. While you can build it using regular op amps, it is probably more widely used in monolithic instrumentation amplifiers (albeit in modified form), difference amplifiers and current shunt monitors.

Fig.10 shows the basic version of this circuit. It provides an extremely useful function; it takes the difference between two voltages, multiplies it by a constant (determined by the resistor values) and then possibly adds a positive or negative offset voltage. However,

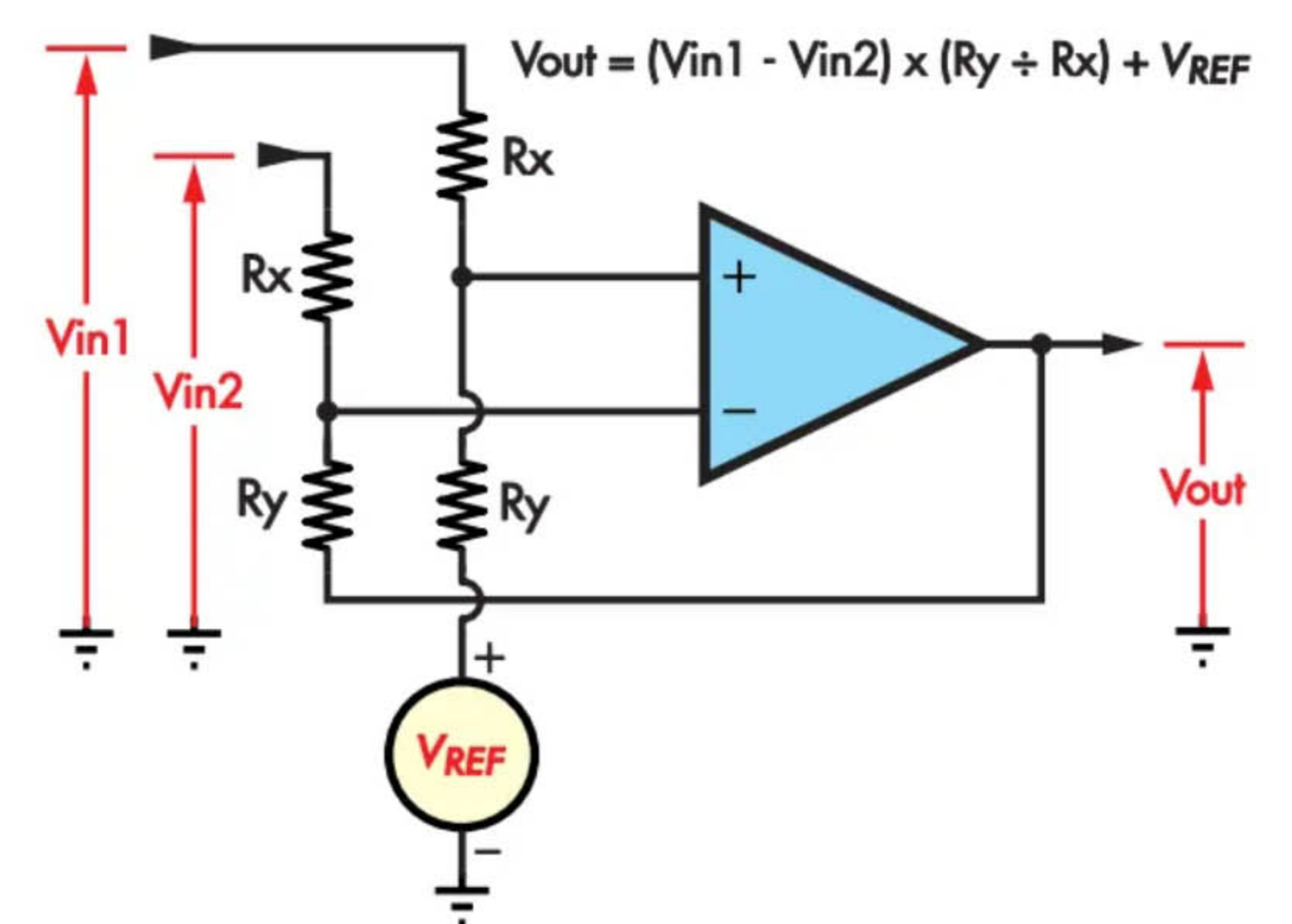


Fig.10: the basic differential amplifier calculates the difference between two voltages, multiplied by a constant, plus an offset. It needs good resistor matching.

er, V_{REF} is often set to 0V so the output voltage is referenced to ground.

This circuit needs precise resistor matching for a good common-mode rejection ratio (CMRR). Even with 0.1% tolerance resistors, a CMRR of more than 60dB is difficult to guarantee. Trimming can give good results, although the procedure can be tricky.

It's generally better to use laser-trimmed monolithic devices like instrumentation amplifiers ('inamps') that can have CMRRs over 100dB.

Most instrumentation amplifiers use a slightly different internal circuit that includes three op amps; besides having a very good CMRR, this has the advantage that the gain can be set using a single external resistor. However, the basic principle is the same.

A difference amplifier is basically an instrumentation amplifier where the input voltages can be well outside (usually above) the device's supply range. A current shunt monitor is a specialised version of an instrumentation amplifier. All are based internally on op amps or similar circuits.

A shunt monitor allows you to place a low-value shunt resistor in the positive

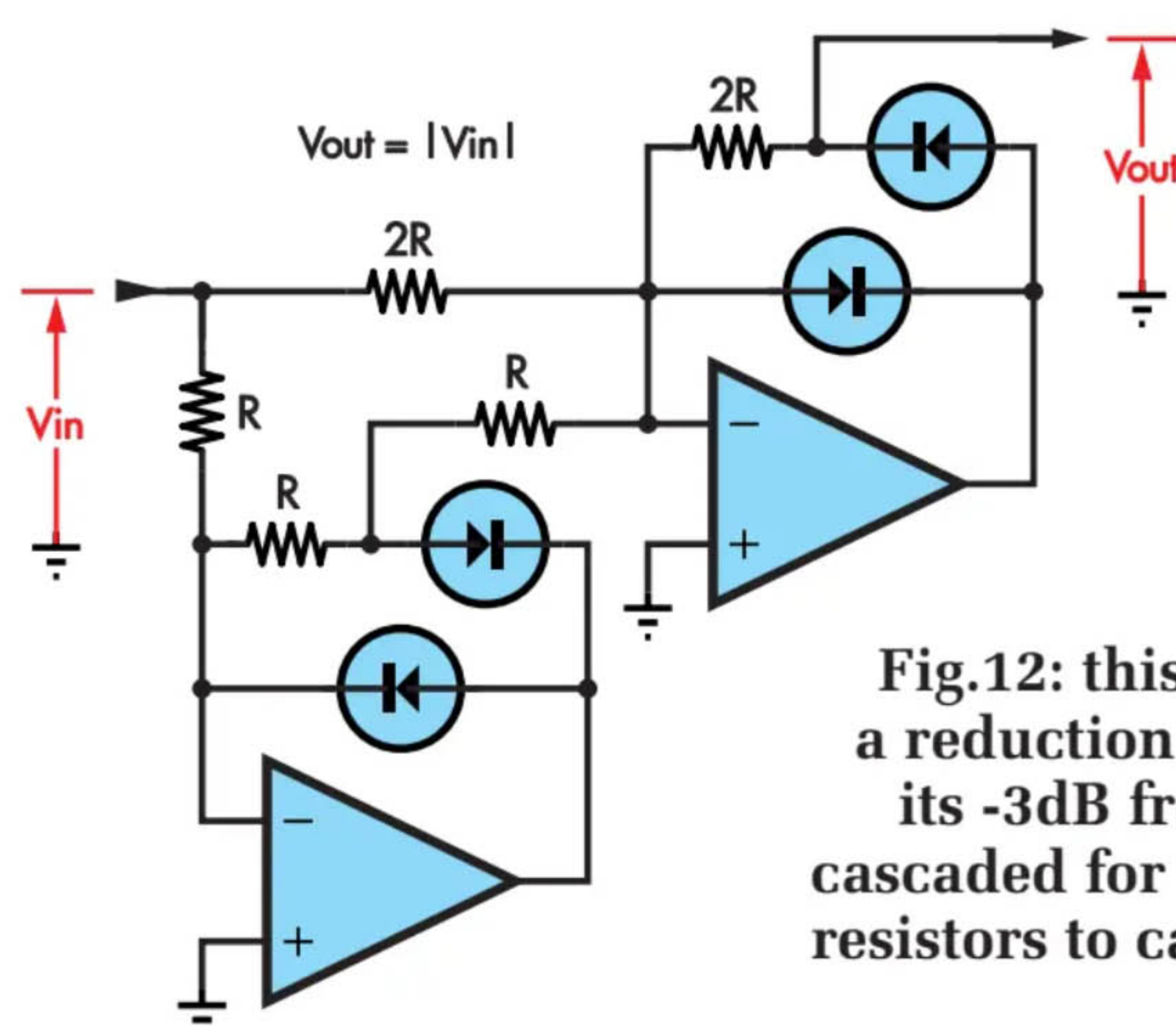
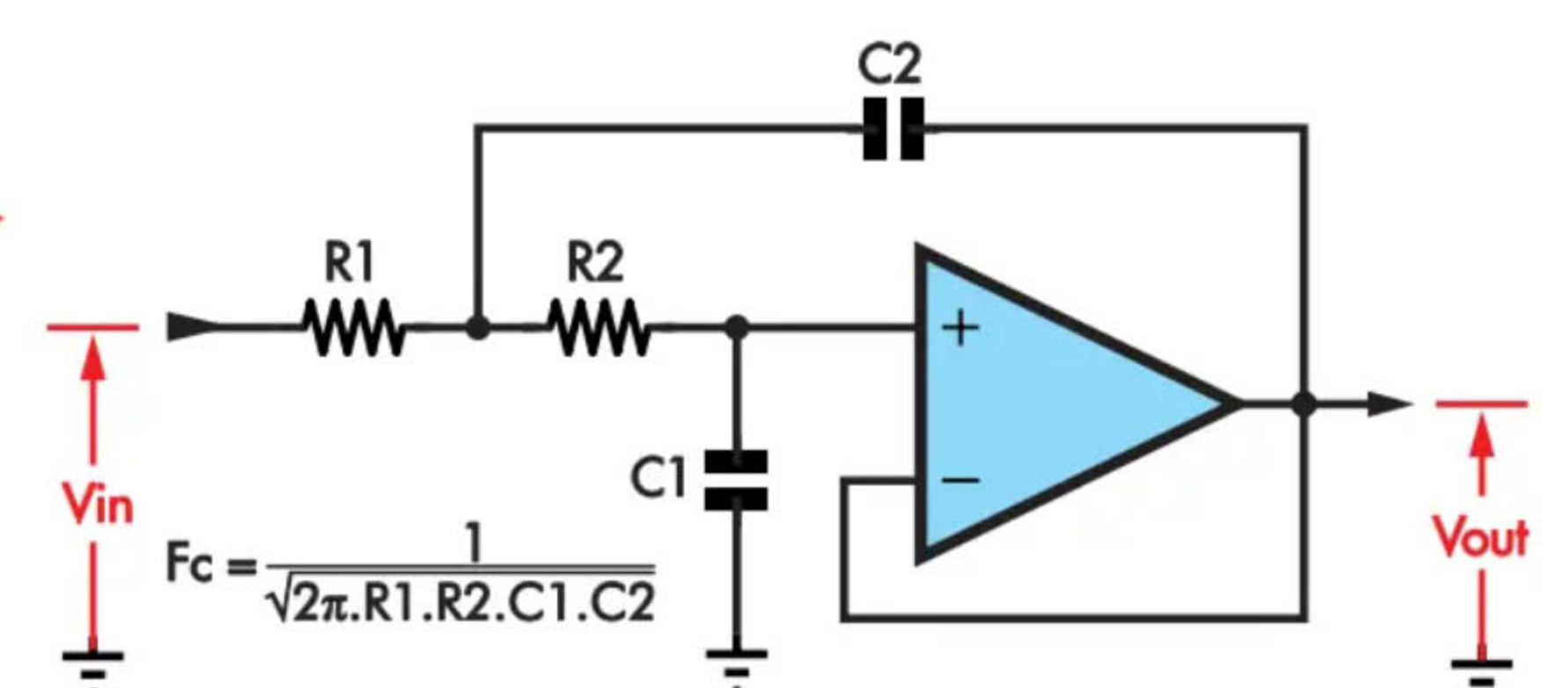
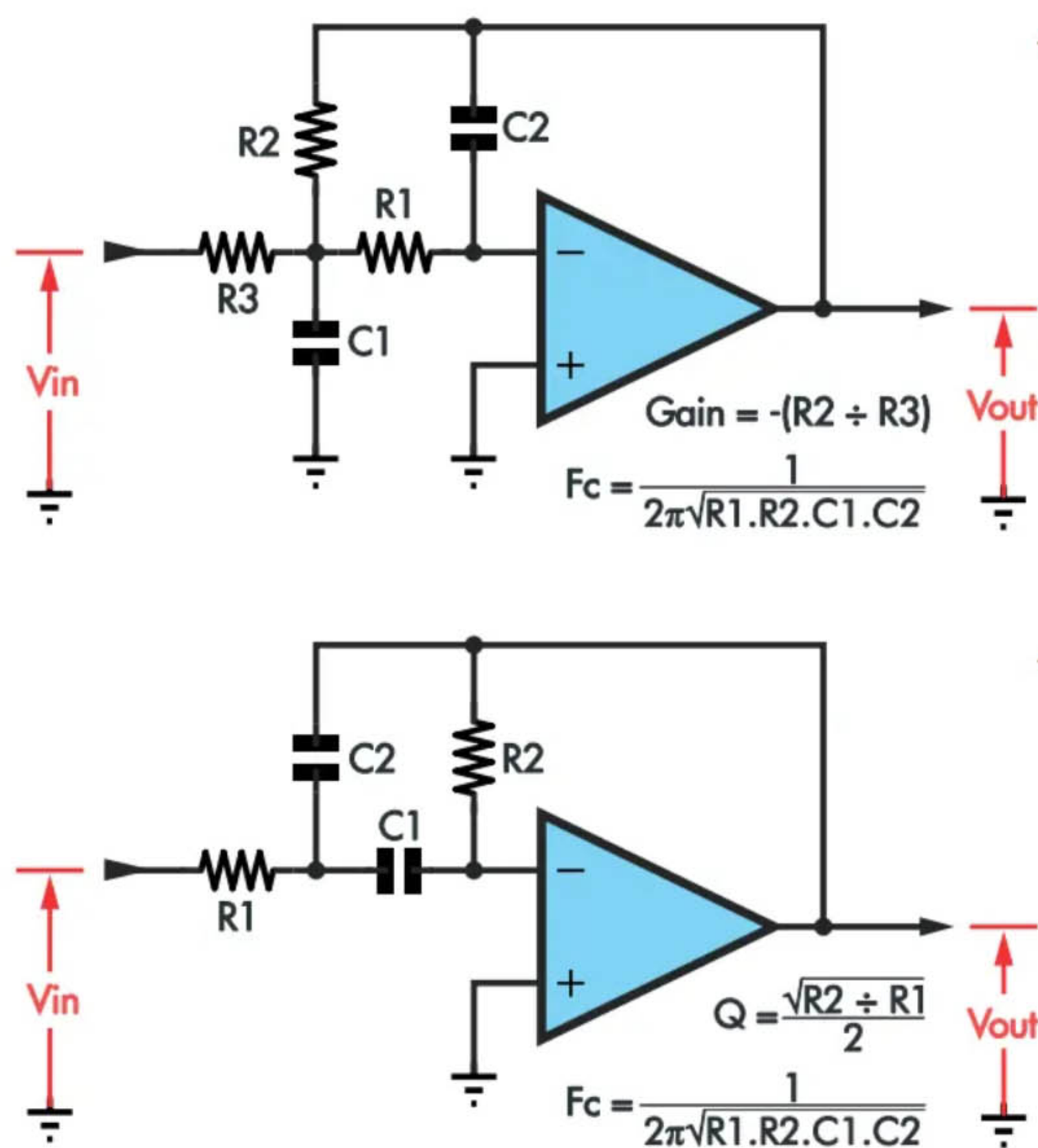


Fig.11: this full-wave rectifier circuit uses op amps to effectively cancel out the forward voltage of the diodes. As a result, for positive voltages at V_{IN} , V_{OUT} tracks very closely (within microvolts, given sufficiently high precision op amps) while for negative voltages at V_{IN} , $V_{OUT} = -V_{IN}$ (again, within microvolts). This is ideal for circuits that need to sense peak signal levels, such as audio clipping meters.

Fig.12: this Sallen-Key low-pass filter provides a reduction in amplitude at -12dB/octave above its -3dB frequency, and multiple stages can be cascaded for an even steeper slope. Changing the resistors to capacitors and capacitors to resistors makes it a high-pass filter instead.



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◀ Fig.13: this multiple feedback filter does the same job as the Sallen-Key filter, but is more effective at higher frequencies. That's important for low-pass filters as otherwise, it can pass signals that the filter is supposed to block. As only one extra resistor is needed, it's a worthwhile upgrade, and the gain can be set without any more resistors (although it does invert the signal).

◀ Fig.14: this active bandpass filter blocks signals outside of a given frequency range, although the slopes are only -6dB/octave. For steeper slopes (eg, -12dB/octave), one of the active low-pass filters described above can be connected in series with a similar high-pass filter.

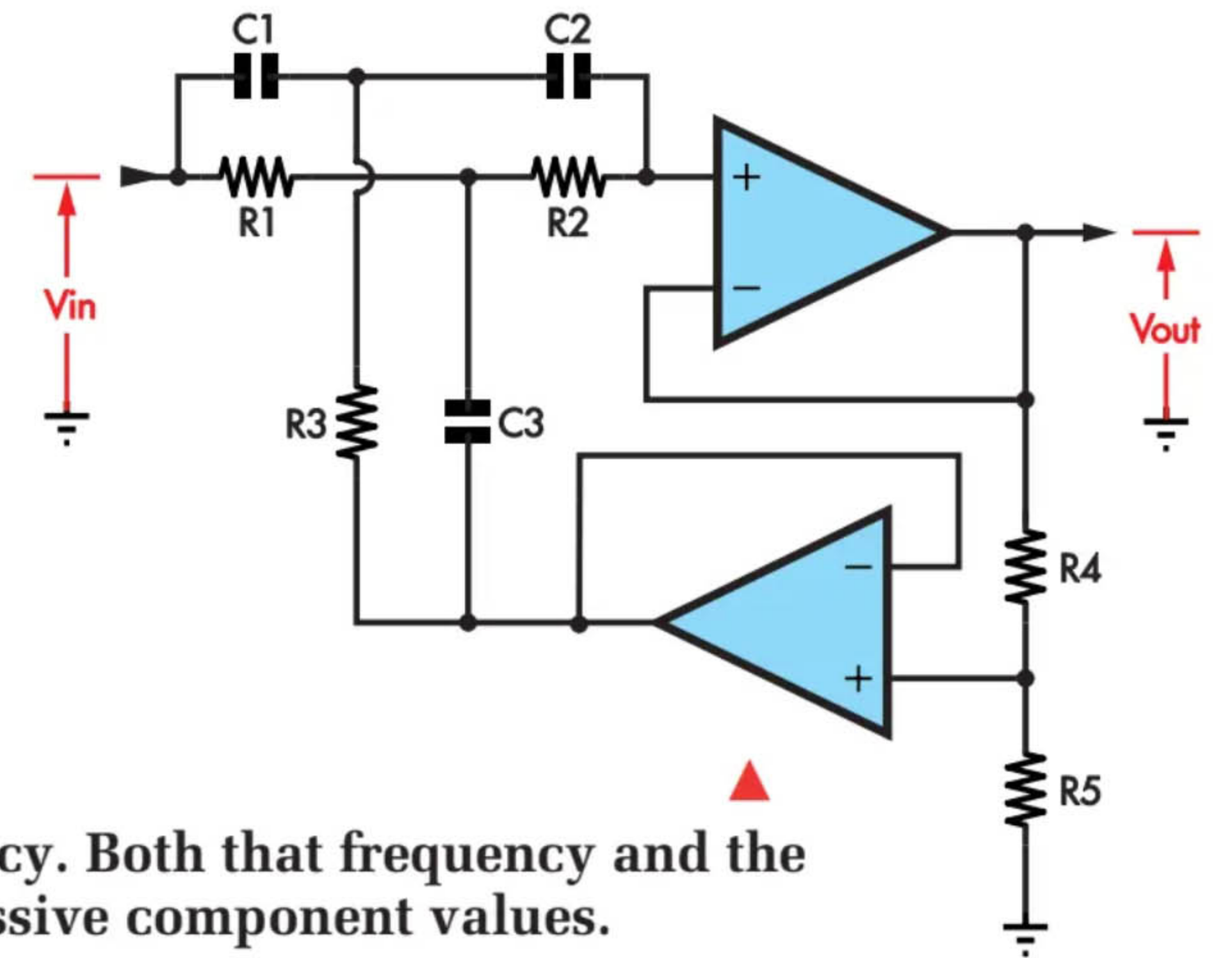


Fig.15: this Twin-T active notch filter attenuates signals at a specific frequency. Both that frequency and the steepness/depth of the notch can be controlled by careful selection of the passive component values.

supply to a section of the circuit, and obtain a ground-referenced voltage to feed to an analog-to-digital converter (ADC) or similar. They have a high CMRR to reject supply ripple.

6) Precision rectifiers

A precision rectifier acts like a diode or bridge rectifier, but without the forward voltage drop. This is important for rectifying low-level signals (too low to forward-bias a diode), or for accurately rectifying AC signals in order to measure their magnitude etc. They are commonly employed in devices like VU meters or AC current monitors.

Fig.11 shows the full-wave version, similar to a bridge rectifier. The half-wave version is basically just one of the op amp/diode/resistor sections.

The op amps reduce the effective forward voltage of the diodes by a factor of their open-loop gain, meaning the ~0.7V drop of a standard silicon diode is effectively less than 1μV for an open-loop gain of around one million.

The resistor values shown result in unity gain. This circuit originally came from National Semiconductor who specified $R = 100k\Omega$, although other values can be used. The values could be changed to give a fixed gain if necessary.

7) Active low-pass filter

The simplest way to implement a low-pass filter with an op amp is to combine a basic RC low-pass filter with a unity-gain buffer. However, a more economical arrangement is the Sallen-Key low-pass filter shown in Fig.12. This has a -12dB/octave slope, compared to -6dB/octave for the RC filter,

using just one op amp. It also allows gain to be applied.

Fig.13 shows a multiple-feedback low-pass filter. This provides precisely the same function as the Sallen-Key filter, but it is less prone to signal feed-through, which means it performs much closer to an ideal filter at frequencies approaching the op amp's bandwidth. The only disadvantage is the use of one more resistor.

To calculate the required resistor and capacitor values for a given cutoff frequency, go to pemag.au/link/aajq

Note that it is possible to build a third-order Sallen-Key active low-pass filter using a single op amp. This will give you an 18dB/octave roll-off with one op amp, 30dB/octave with two etc. This is shown at pemag.au/link/ab8v

8) Active high-pass filter

To convert the low-pass filters shown in Figs.12 & 13 into high-pass filters, simply transpose the resistors and capacitors. As with the low-pass filters, these will provide a 12dB/octave slope per op amp.

For both the low-pass and high-pass filters, by adjusting the resistances and capacitances, it is possible to design filters with characteristics other than Butterworth. Butterworth has minimal (essentially no) ripple in the passband, but different filter types such as Chebyshev trade off increased passband ripple for a steeper roll-off beyond it.

To calculate the required component values, see pemag.au/link/ab8w

9) Active bandpass filter

A second-order bandpass filter can be created by combining active second-

order low-pass and high-pass filters. Alternatively, you can use the configuration shown in Fig.14, where a single op amp can act as a first-order band-pass filter with adjustable gain and a Q of up to 25. This configuration inverts the signal phase; however, if chaining multiple filters, it can be re-inverted by another stage.

10) Active notch filter

Fig.15 shows a "Twin-T" active notch filter. One interesting aspect of this design is that the Q, and thus the depth of the notch, changes based on the resistor and capacitor values selected. See the online calculator at pemag.au/link/ab8x

11) Gyrator

Fig.16 shows a 'gyrator', an active element that behaves similarly to an ideal inductor at low current values. It does this by using the op amp's negative feedback to effectively invert the behaviour of capacitor C.

This can be useful in circuits like

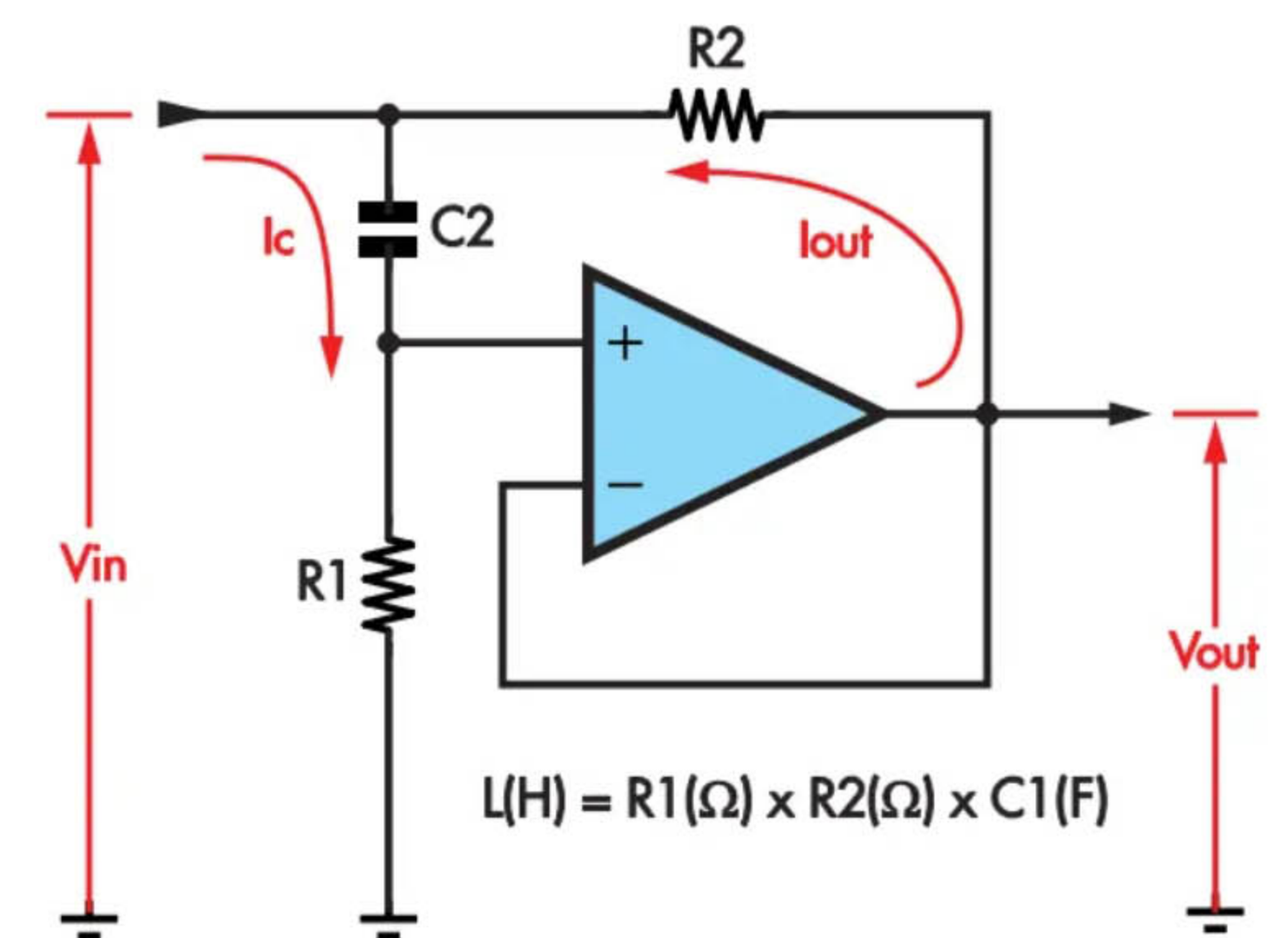


Fig.16: the gyrator is a particularly clever circuit. It uses negative feedback to make a capacitor behave like an inductor. It is superior to an actual inductor in many signal processing applications.

graphic equalisers, where resonant (LC) elements are needed with accurate resonance frequencies, low distortion and small size. Inductor tolerances are typically much wider than capacitors, and high-value inductors can be very bulky, so in signal-processing circuits, the gyrator is almost always better than a resonant circuit based on an actual inductor.

12) Baxandall active filter

Fig.17 shows a basic version of the widely-used Baxandall active tone control. It has many good properties, such as the ability to have as many or as few bands as you want, with no interaction between the controls, and no special requirements for the potentiometers. This one shows bass and treble pots, but one or two midrange pots can easily be added.

Fig.18 is the Baxandall active volume control. The traditional volume control method is a logarithmic potentiometer, but dual versions usually have poor tracking at the low end, so they are not great for stereo circuits.

The Baxandall active circuit provides logarithmic-like control with a linear potentiometer for superior tracking. It can also offer significantly better noise performance as the pot adjusts the gain over a wide range, from zero up to many times (as set by the fixed resistors).

13) Audio amplifiers

Fig.19 is a simplified version of the circuit from the SC200 audio amplifier. It is essentially a high-power op amp with large output transistors that can source and sink plenty of current (and that are well heatsinked).

Most Class-A, Class-AB and similar amplifiers are variations on this

theme. Even Class-D amplifiers typically use some form of negative feedback to avoid gross distortion.

14) Other uses for op amps

An op amp can be used as a basic comparator by operating it in open-loop mode, or with positive feedback (hysteresis). A comparator IC is essentially just an op amp with the frequency compensation component(s) removed for a faster swing at the output.

An op amp can also be used to build an 'integrator' or 'differentiator'. An integrator produces an output ramp proportional to its input voltage, while a differentiator produces an output voltage that's proportional to its input ramp (rate of change).

A log amp takes the exponential nature of a bipolar transistor and turns it on its head using negative feedback to provide a logarithmic transfer function. As a result, its output voltage is a constant multiple of the natural logarithm of its input voltage.

This can be used as the basis of a multiplier circuit; by taking the natural $\log_e(x)$ of several voltages, summing or averaging them, then exponentiating the result, the output voltage is the product of the input voltages.

Other mathematical functions can be applied to voltages by an op amp, including addition, subtraction, division and inverse logarithm (the exponentiation mentioned above).

Op amps can also be used to build controlled current sources/sinks, including constant loads, by combining op amps with large transistors that can handle lots of power with sufficient heatsinking.

The generalised impedance con-

verter uses two op amps to present a load impedance proportional to another impedance. The ratio can be set using fixed or variable resistors (or even other impedances!).

Many op amps are designed to drive relatively low load impedances, such as 600Ω. These work quite well as basic headphone drivers, with relatively low distortion figures driving typical headphone loads, even as low as 16Ω. They can't supply a tremendous amount of power, but enough for most headphones to deliver decent volume, using one low-cost IC.

An op amp can also be used as an error amplifier in feedback control. For example, to adjust the drive to a motor to maintain a constant speed despite a varying load.

An op amp can form the basis of various oscillators, to generate waveforms at fixed or variable frequencies; primarily sinewaves, but also triangle waves or sawtooth waveforms.

An op amp (especially a CMOS type) can be used as a high input-impedance buffer amplifier or guard ring for monitoring sensors that cannot handle any loading, such as narrow-band oxygen sensors and pH sensors. CMOS op amps can have input impedances in the terohms range (more than one trillion ohms)!

CMOS op amp buffers can also be combined with analog switches and low-leakage capacitors to form sample-and-hold circuits, often used for sampling voltages over small time windows to feed an ADC or similar.

Signal swing limitations

For a very long time, the signals at the inputs and outputs of an op amp

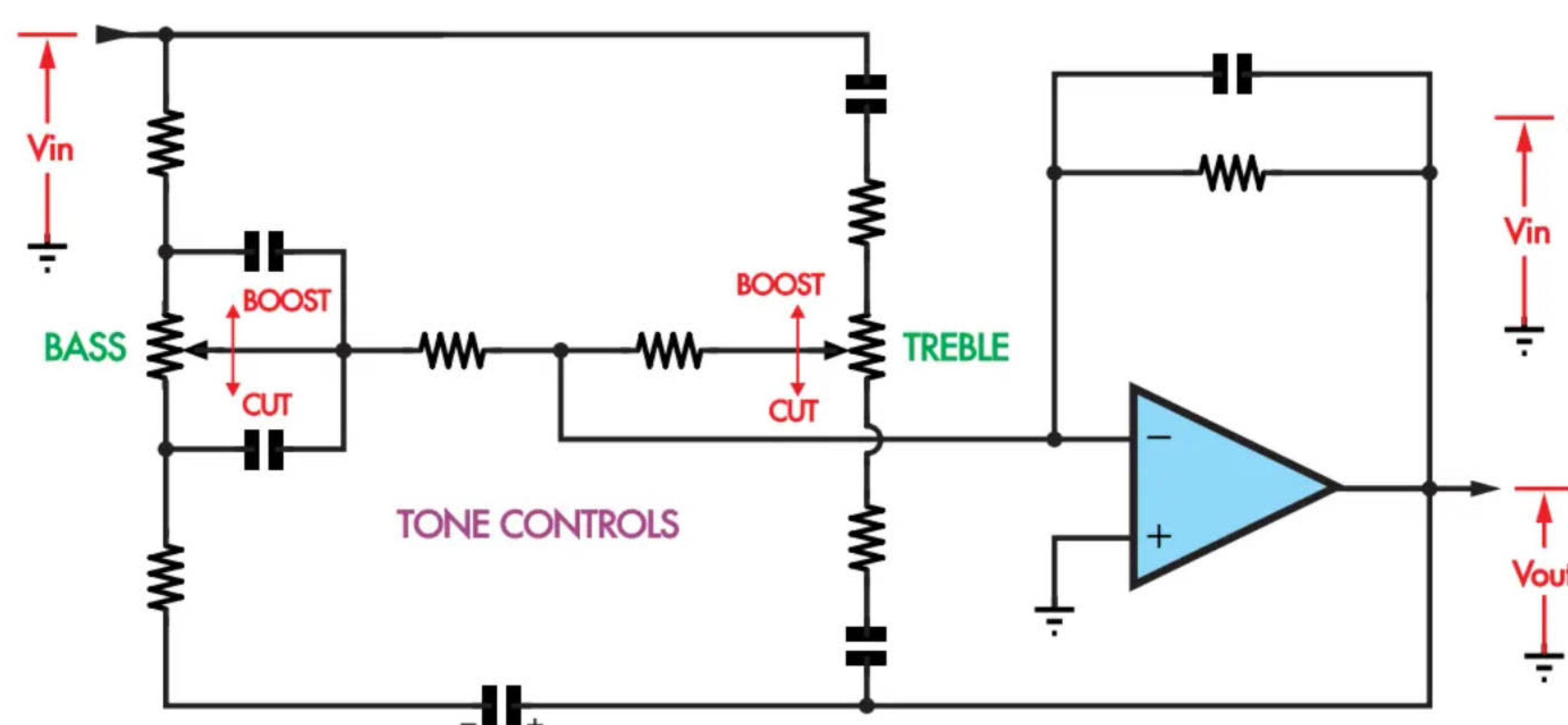


Fig.17: the Baxandall tone control was initially designed with a valve or transistor as the active element, but it works even better with an op amp. It is elegant and expandable, with virtually no interaction between the stages (in this case, two: bass and treble adjustments). No matter how many bands it has, only one op amp is required per channel (ie, two for stereo).

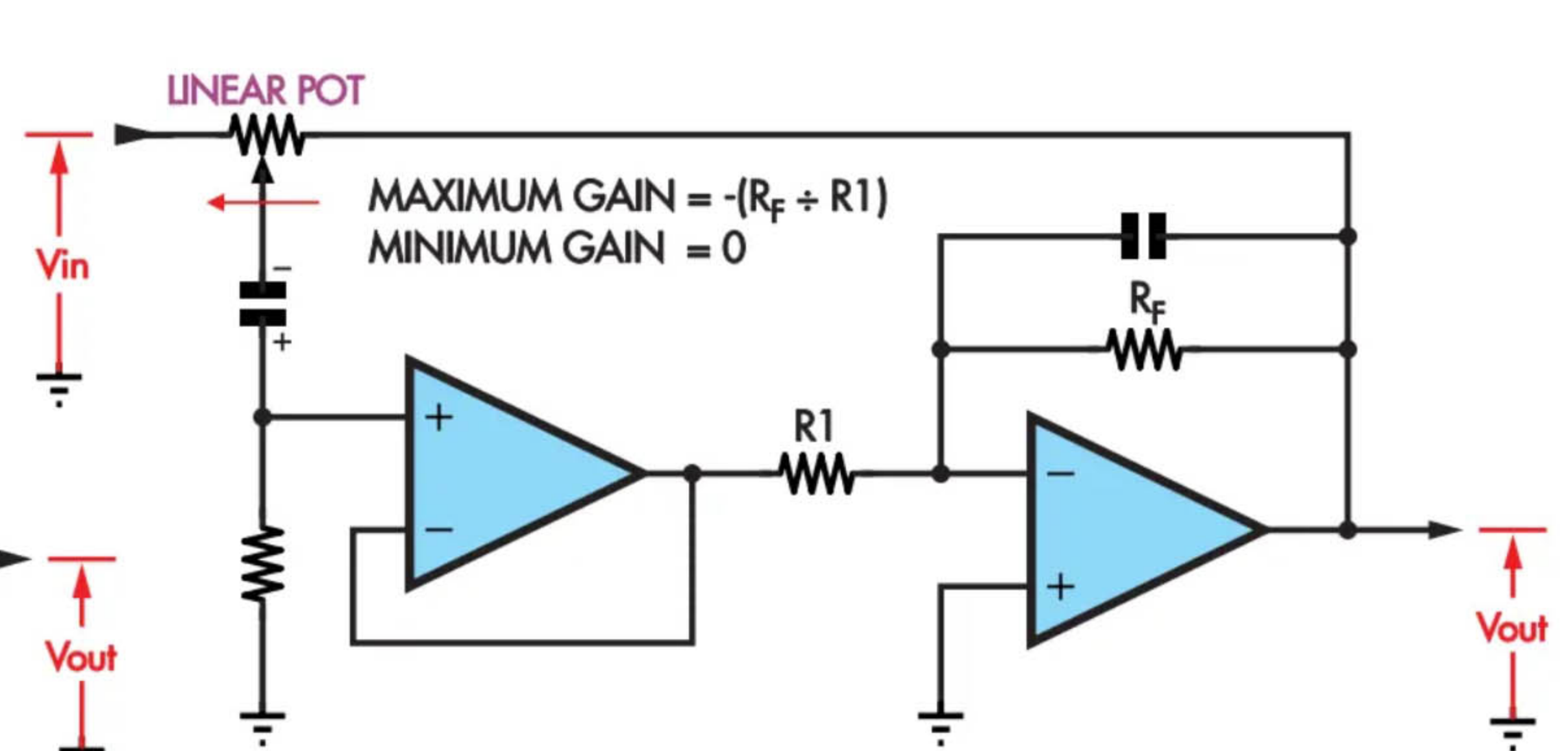


Fig.18: the Baxandall volume control also places the potentiometer in the negative feedback loop. This gives exponential gain control with a linear potentiometer and a wide range of gain settings with a reasonably constant noise level.

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could only have a considerably smaller swing than the supply range of the op amp. For example, if you had an op amp running from 12V, the inputs and outputs might be limited to a range of 3-9V. Or, with a dual supply like $\pm 15\text{V}$, you might be limited to a signal swing of $\pm 12\text{V}$.

That's because the op amp's internal circuitry needs some voltage 'headroom' to operate.

But more recently, single-supply and rail-to-rail output op amps started to become available. Single-supply op amps typically allow the inputs and outputs to go down to the negative rail (eg, 0V). So a single-supply op amp running from 12V can handle signals of say 0-9V.

Rail-to-rail output op amps generally have the same input limitations as standard op amps, but their output can swing over virtually the entire supply range. This is especially useful when applying gain to AC signals, as in that case, the input swing will never reach the rails anyway (at least, not without 'saturating' the op amp).

These days, rail-to-rail input/output

(RRIO) op amps are very common. Some can even run down to very low supply voltages, like 1.8V! These op amps essentially remove the above limitations, with input and output signals that can range anywhere between the supply rails.

Some will even handle input signals outside the rails, although usually only in one direction (eg, positive) and by a limited number of volts.

Note that RRIO op amps sometimes compromise performance in other ways, such as having higher noise or distortion, or just costing more than 'regular' op amps.

Multiple op amps

As op amps became cheaper and more versatile, dual and quad op amps became popular. These save money and space; a quad op amp IC often costs less than twice what a single one does, and only requires two power tracks to be routed and one bypass capacitor. Most dual (8-pin) and quad (14-pin) op amp ICs use the same pinout so they can be interchanged.

Single op amps are not quite so in-

terchangeable, as these usually come in an 8-pin package. After accounting for the two supply rails, two inputs and one output, the remaining three pins can be used for trimming/balancing, external compensation capacitors or various other functions. Some are interchangeable (even if they don't have exactly the same features), but not all.

These days, single op amps are also available in tiny 5-pin SMD packages for where space is at a premium.

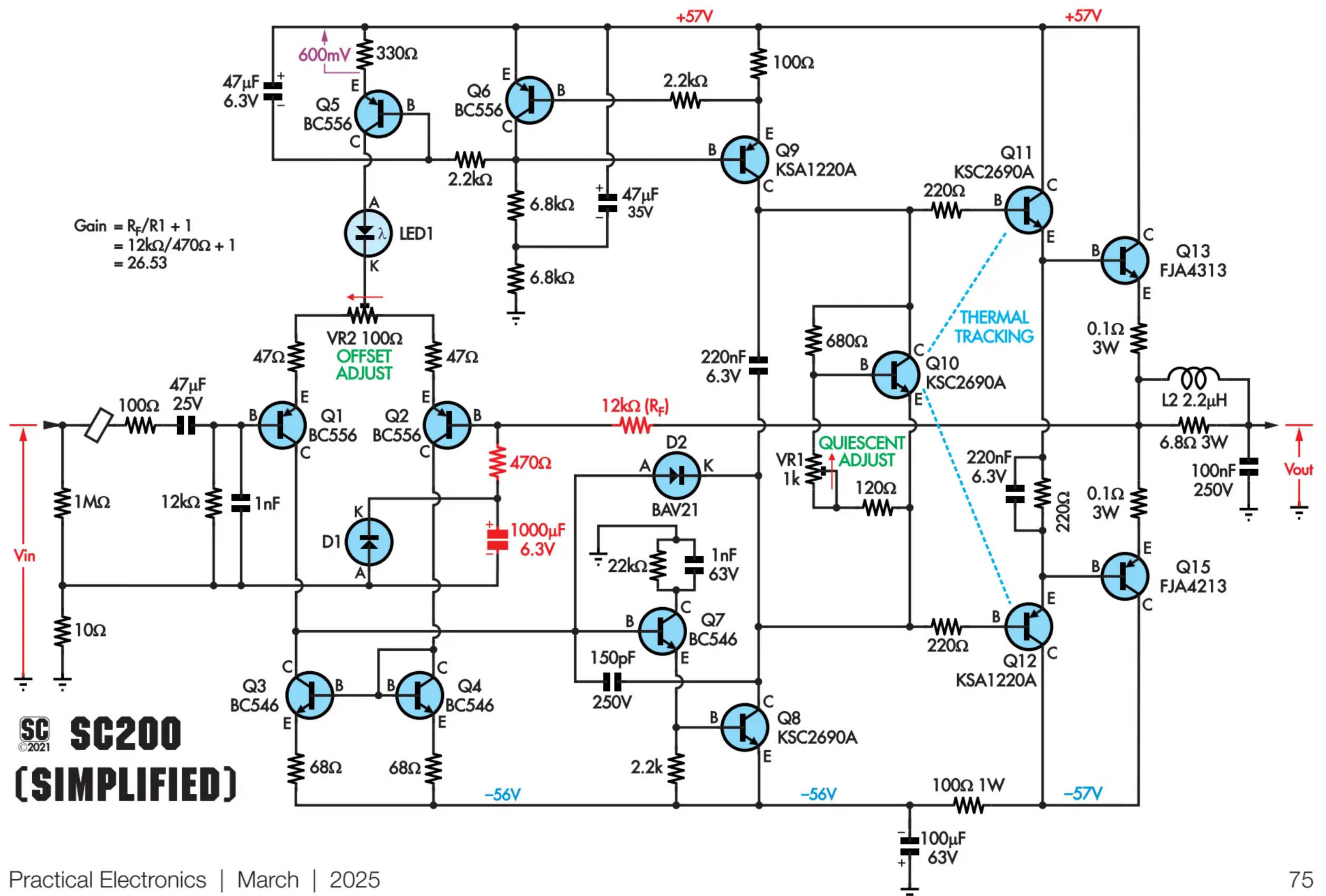
Conclusion

The op amp is an incredibly flexible device, available these days at very low cost and in a vast range of different versions, optimised for different tasks. While it is possible to process analog signals without op amps, generally, the results will be worse. So most analog designers make extensive use of op amps in their circuitry.

They are an essential building brick that most designers would have difficulty doing without. We have Harold S. Black to thank for making our lives a lot easier!

PE

Fig.19: a slightly simplified version of our SC200 power amplifier circuit. It's essentially a big op amp; transistors Q1 & Q2 are the differential input pair (the inputs are at their bases), Q8 is the voltage amplification stage, Q11 & Q12 are the output drivers and Q13 and Q15 are the power output transistors. The components highlighted in red form the negative feedback path, from the output at the emitter resistors of Q13 & Q15 back to the base of Q2, which is the inverting input.



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PCBs for most recent *PE/EPE* constructional projects are available. From the July 2013 issue onwards, PCBs with eight-digit codes have silk screen overlays and, where applicable, are double-sided, have plated-through holes, and solder mask. They are similar to photos in the project articles. Earlier PCBs are likely to be more basic and may not include silk screen overlay, be single-sided, lack plated-through holes and solder mask.

Always check price and availability in the latest issue or online. A large number of older boards are listed for ordering on our website.

In most cases we do not supply kits or components for our projects. For older projects it is important to check the availability of all components before purchasing PCBs.

Magazine back issues are available – see <https://www.electronpublishing.com/product-category/pe-back-issues-print-and-digital/>

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			PE Theremin Amplifier	AO-1220-01	£8.95
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Compact Frequency Divider kit	https://siliconchip.au/Shop/20/6881	AUD \$48.00 (about £24) + P&P
Mains Power-Up Sequencer programmed PIC	https://siliconchip.au/Shop/20/6901	AUD \$12.00 (about £6) + P&P
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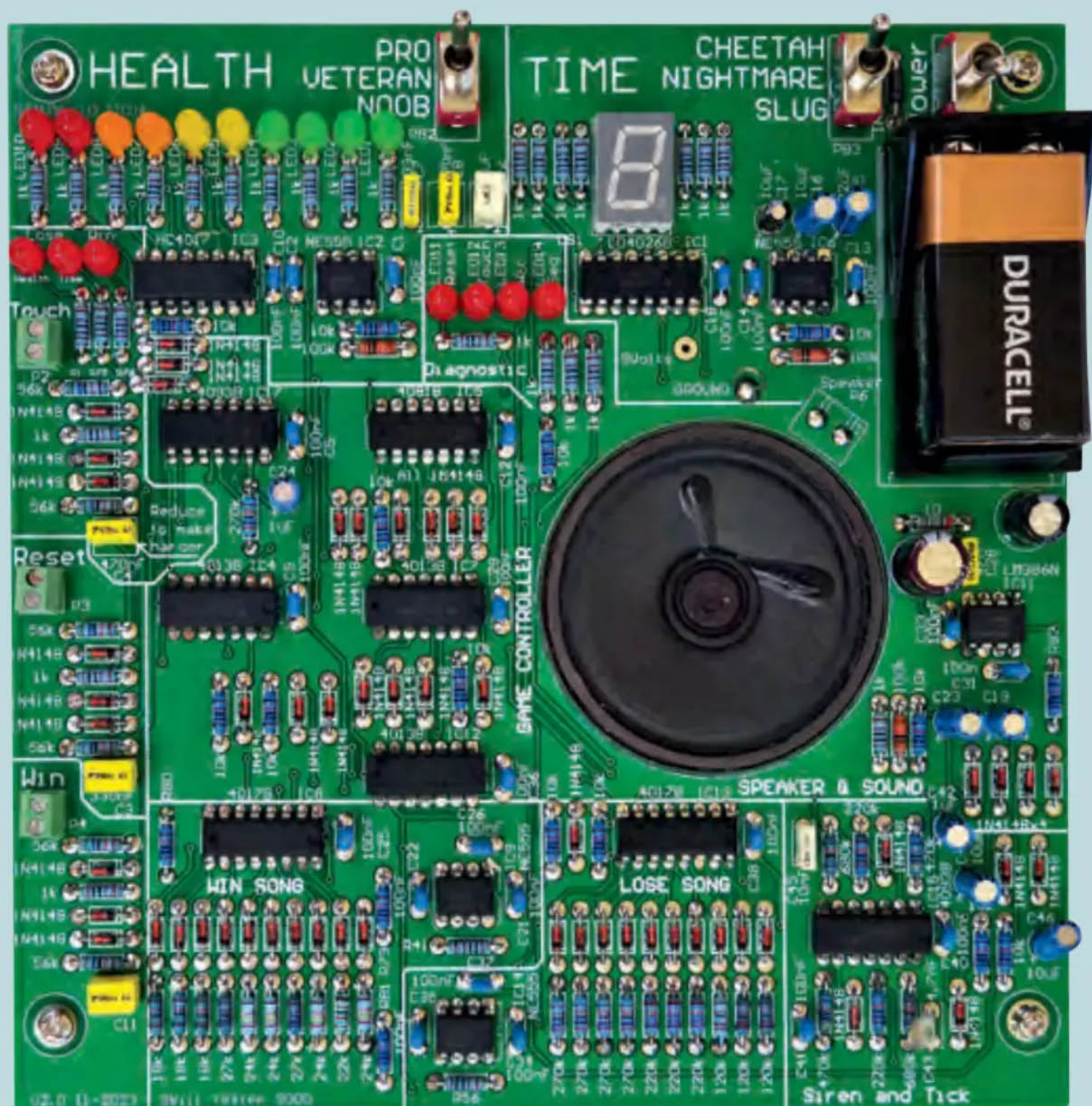
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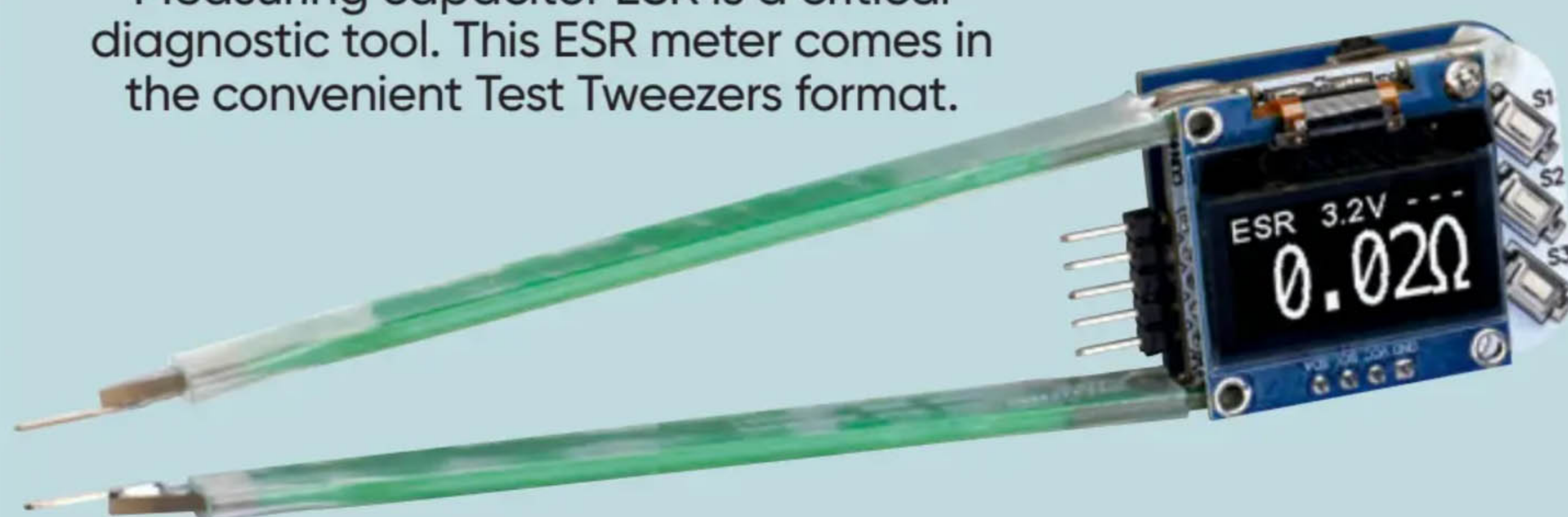
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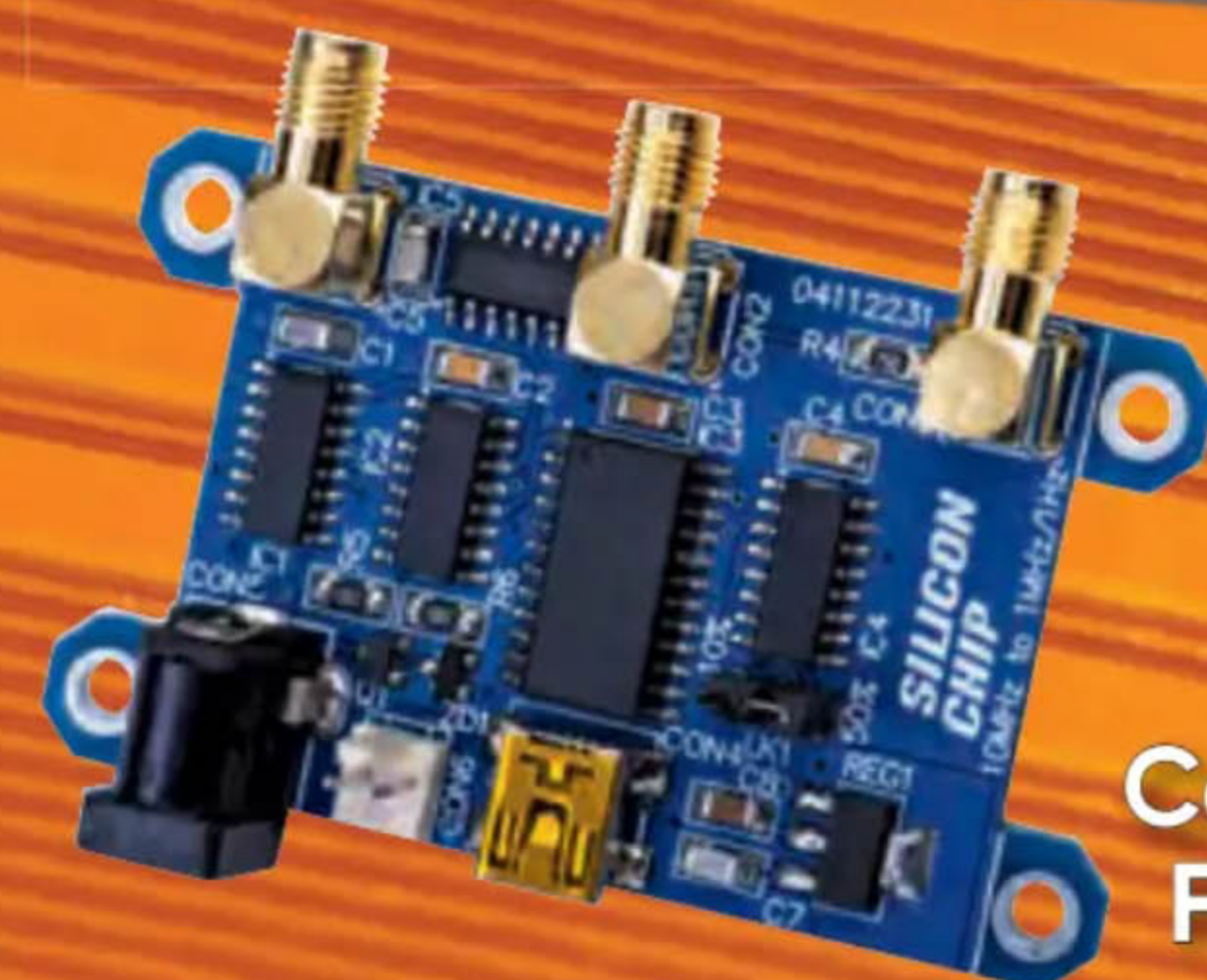
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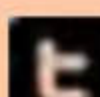
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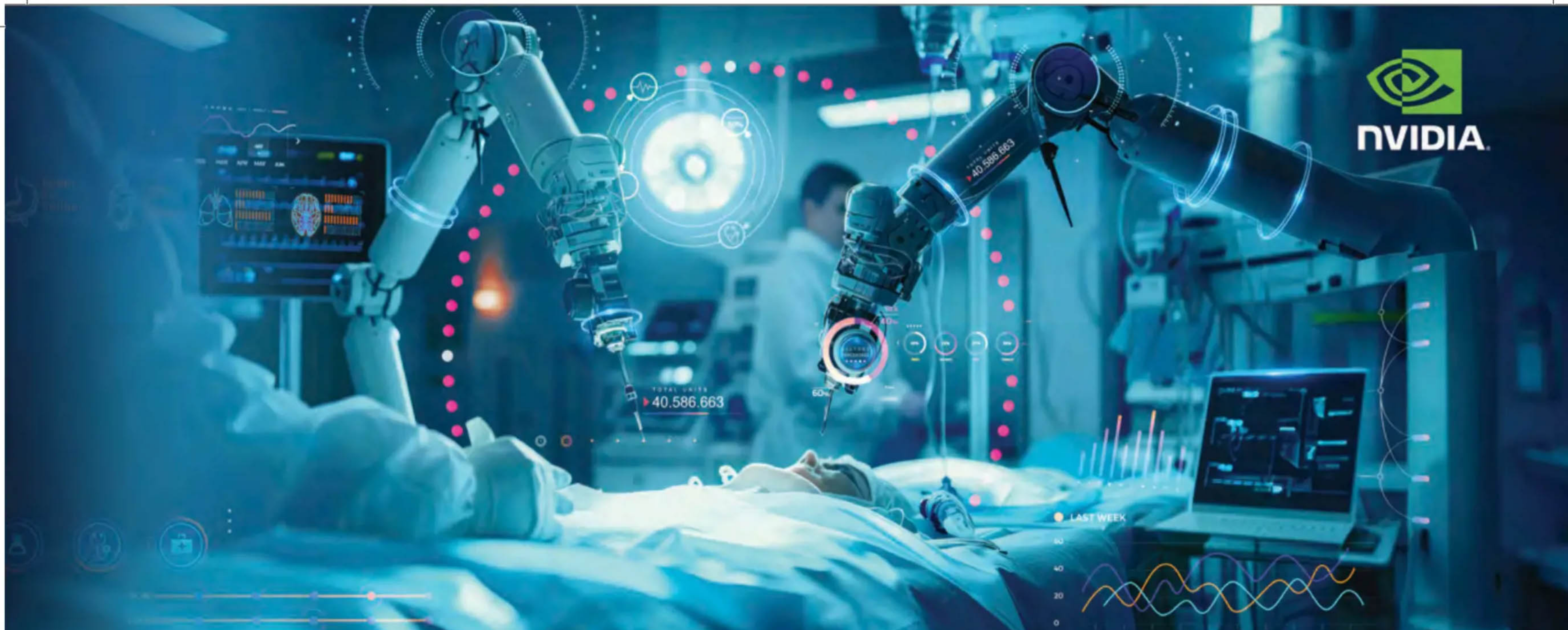
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